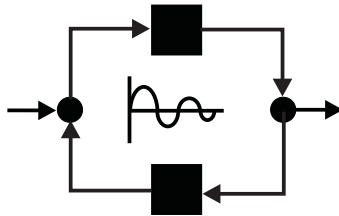




**The International Congress for global
Science and Technology**



**ICGST International Journal on Automatic
Control & System Engineering
(ACSE)**

**Volume (8), Issue (III)
January, 2009**

**www.icgst.com
www.icgst-amc.com
www.icgst-ees.com**

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ACSE Journal
ISSN: Print 1687-4811
ISSN Online 1687-482X
ISSN CD-ROM 1687-4838
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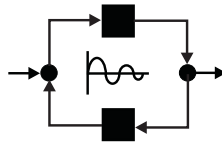
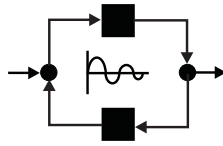


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ICGST International Journal on Automatic Control & System Engineering - (ACSE)

**A publication of the International Congress for global Science and Technology -
(ICGST)**

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SENSORLESS NONLINEAR ADAPTIVE BACKSTEPPING CONTROL OF INDUCTION MOTOR

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Abstract

A Sensorless Backstepping technique combined with a field orientation scheme has been developed for the control of induction motor to achieve rotor angular speed and rotor flux amplitude tracking objectives. An accurate knowledge of the rotor speed is the key factor in obtaining a high-performance and high-efficiency induction-motor drive. For that, a Model reference adaptive system (MRAS) rotor-speed estimator has been incorporated for which stability and robustness are assured. Simulation results are presented to validate the effectiveness of the overall proposed sensorless control scheme under the extreme conditions.

Keywords: Induction motor, Backstepping design, Vector control, MRAS speed identification.

NOMENCLATURE

R_1, R_2	stator, rotor resistance
i_1, i_2	stator, rotor current
λ_1, λ_2	stator, rotor flux linkage
λ_{2d}	amplitude of rotor flux linkage
u_1, u_2	stator, rotor voltage input
ω	rotor angular speed
ρ	angle between the rotor flux linkages
n_p	number of pole pairs

L_1, L_2, L_m stator, rotor and mutual inductance

T_1, T_2 stator, rotor time constant

J, T_L inertia, load torque

$(\cdot)_d, (\cdot)_q$ in (d, q) frame

$(\cdot)_a, (\cdot)_b$ in (a, b) frame

$(\cdot)$ estimate of $(\cdot)$

$$\sigma = 1 - \frac{L_m^2}{L_1 L_2}, \quad \alpha = \frac{1}{L_2}, \quad \beta = \frac{L_m}{\sigma L_1 L_2}, \quad \eta_1 = \frac{R_1}{\sigma L_1}$$

$$\eta_2 = \frac{L_m^2}{\sigma L_1 L_2^2}, \quad \mu = \frac{3n_p L_m}{2J L_2}, \quad p = d/dt$$

1. Introduction

Induction motors have been widely used in high performance alternating current (ac) machine drives, requiring speed information. Introducing shaft speed sensor decreases system reliability. Different solutions for speed sensorless control of the induction motor (IM) have been proposed [1, 2, 3]. Among them, MRAS based techniques have been proven to be one of the best methods being proposed by the researchers, due to its design and implementation simplicity and high performance capability[4].

In the last two decades, many modified nonlinear state feedback schemes such as input-output feedback linearization [5], passivity-based control [6, 7] and Sliding-Mode (SM) control [8] have been applied to the IM drive.

Specially in the last few years, in the field of adaptive and robust control, there has been a tremendous amount of activity on a special control scheme known as "Backstepping" [9, 10, 11]. With adaptive Backstepping design technique researchers were able to successfully design controllers which achieved global stabilization in the presence of uncertain parameters[12, 13]. Using this design methodology the construction of both the feedback control laws and associated Lyapunov functions is systematic. Strong properties of global or local stability are built into the nonlinear system with uncertain parameters [14].

In this paper, we have integrated a powerful tool of control (Backstepping) with a powerful tool of speed estimation (MRAS). To validate the effectiveness of this integration, the overall scheme is subjected to the extreme variation of speed reference and the benchmark test.

The rest of this paper is organized as follows: in section-2, we present the IM model used and the Backstepping control design, in section-3, we present the rotor flux estimator, in section-4 the MRAS speed estimator is detailed and in section-5 the stability of the overall scheme is discussed.



2. Model Description and Control Design

Assuming linear magnetic circuits and balanced three phase windings, the fifth-order nonlinear model of IM, expressed in the stator frame is [15]:

$$\begin{aligned}
 \frac{d\omega}{dt} &= \frac{3n_p L_m}{2JL_2} (\lambda_{2a} i_{1b} - \lambda_{2b} i_{1a}) - \frac{T_L}{J} \\
 \frac{d\lambda_{2a}}{dt} &= -\frac{R_2}{L_2} \lambda_{2a} - n_p \omega \lambda_{2b} + \frac{R_2}{L_2} L_m i_{1a} \\
 \frac{d\lambda_{2b}}{dt} &= -\frac{R_2}{L_2} \lambda_{2b} + n_p \omega \lambda_{2a} + \frac{R_2}{L_2} L_m i_{1b} \\
 \frac{di_{1a}}{dt} &= \frac{L_m R_2}{\sigma L_1 L_2^2} \lambda_{2a} + \frac{n_p L_m}{\sigma L_1 L_2} \omega \lambda_{2b} - \frac{L_m^2 R_2 + L_2^2 R_1}{\sigma L_1 L_2^2} I_{1a} \\
 &\quad + \frac{1}{\sigma L_1} u_{1a} \\
 \frac{di_{1b}}{dt} &= \frac{L_m R_2}{\sigma L_1 L_2^2} \lambda_{2b} - \frac{n_p L_m}{\sigma L_1 L_2} \omega \lambda_{2a} - \frac{L_m^2 R_2 + L_2^2 R_1}{\sigma L_1 L_2^2} I_{1b} \\
 &\quad + \frac{1}{\sigma L_1} u_{1b}
 \end{aligned} \tag{1}$$

We can see that (1) is highly coupled, multivariable and nonlinear system. It is very difficult to control the IM directly based on this model. State transformation to simplify the system representation is required. A well-known method to this end is the transformation of the field orientation principle. It involves basically a change of the representations of the state vector $(i_{1a}, i_{1b}, \lambda_{2a}, \lambda_{2b})$ in the fixed stator frame (a, b) into a new state vector in a frame (d, q) which rotates along with the flux vector $(\lambda_{2a}, \lambda_{2b})$.

Mathematically, the field oriented transformation can be described as:

$$\begin{aligned}
 i_{1d} &= \frac{\lambda_{2a} i_{1a} + \lambda_{2b} i_{1b}}{\sqrt{\lambda_{2a}^2 + \lambda_{2b}^2}}, \quad i_{1q} = \frac{\lambda_{2a} i_{1b} - \lambda_{2b} i_{1a}}{\sqrt{\lambda_{2a}^2 + \lambda_{2b}^2}} \\
 \lambda_{2d} &= \sqrt{\lambda_{2a}^2 + \lambda_{2b}^2}, \quad \lambda_{2q} = 0, \quad \rho = \arctan\left(\frac{\lambda_{2b}}{\lambda_{2a}}\right)
 \end{aligned} \tag{2}$$

Where much simplification is gained by the fact that $\lambda_{2q} = 0$.

Using this transformation and the notations in the Nomenclature, the state equations (1) can be rewritten in the new state variables as:

$$\begin{aligned}
 \frac{d\omega}{dt} &= \mu \lambda_{2d} i_{1q} - \frac{T_L}{J} \\
 \frac{di_{1q}}{dt} &= -\eta_1 i_{1q} - \beta n_p \omega \lambda_{2d} - n_p \omega i_{1d} - R_2 \left(\eta_2 I_{1q} + \alpha L_m \frac{i_{1q} i_{1d}}{\lambda_{2d}} \right) \\
 &\quad + \frac{1}{\sigma L_1} u_{1q} \\
 \frac{d\lambda_{2d}}{dt} &= -\alpha R_2 \lambda_{2d} + \alpha L_m R_2 i_{1d} \\
 \frac{di_{1d}}{dt} &= -\eta_1 i_{1d} + n_p \omega i_{1q} + R_2 \left(-\eta_2 I_{1d} + \alpha \beta \lambda_{2d} + \alpha L_m \frac{i_{1q}^2}{\lambda_{2d}} \right) \\
 &\quad + \frac{1}{\sigma L_1} u_{1d} \\
 \frac{d\rho}{dt} &= n_p \omega + \alpha L_m R_2 \frac{I_{1q}}{\lambda_{2d}}
 \end{aligned} \tag{3}$$

From this structure, the Backstepping control mechanisms for the rotor angular speed regulation and the flux generation can be better applied to replace the traditional nonlinear feedback plus (proportional-integral) PI control of the field oriented control technique for better performance.

The basic idea of the Backstepping design is the use of the so-called “virtual control” to systematically decompose a complex nonlinear control design problem into simpler, smaller ones. Roughly speaking, Backstepping design is divided into various design steps. In each step we essentially deal with an easier, single-input-single-output design problem, and each step provides a reference for the next design step. The overall stability and performance are achieved by a Lyapunov function for the whole system [16].

Now we turn to the Backstepping design steps. First, for the rotor angular speed, and rotor flux amplitude tracking objectives, define the tracking errors as

$$\begin{aligned}
 e_1 &= \omega_{ref} - \omega \\
 e_3 &= \lambda_{ref} - \lambda_{2d}
 \end{aligned} \tag{4}$$

Then the error dynamical equations are

$$\begin{aligned}
 \dot{e}_1 &= \dot{\omega}_{ref} - \mu \lambda_{2d} i_{1q} + \frac{T_L}{J} \\
 \dot{e}_3 &= \dot{\lambda}_{ref} + \alpha R_2 \lambda_{2d} - \alpha L_m R_2 I_{1d}
 \end{aligned} \tag{5}$$

Since our objectives require that the two errors converge to zero, we could satisfy the objectives by viewing $\lambda_{2d} i_{1q}$ and $R_2 i_{1d}$ as “virtual control” in the above equations and use them to control e_1, e_2 . Certainly, $\lambda_{2d} i_{1q}$ is only a constant from the torque generated by the motor, and we include R_2 in the virtual control $R_2 i_{1d}$. We use the Lyapunov function

$$V = \frac{1}{2} e_1^2 + \frac{1}{2} e_3^2 \tag{6}$$



To derive stabilizing feedback virtual controls $\lambda_{2d}i_{1q}$ and R_2i_{1d} as follows. The derivatives of V along the error equations (5) is

$$\begin{aligned}\dot{V} &= e_1\dot{e}_1 + e_3\dot{e}_3 \\ &= e_1\left[\dot{\omega}_{ref} - \mu\lambda_{2d}i_{1q} + \frac{T_L}{J}\right] \\ &\quad + e_3\left[\dot{\lambda}_{ref} + \alpha R_2\lambda_{2d} - \alpha L_m R_2i_{1d}\right] \\ &= -k_1e_1^2 - k_3e_3^2 + e_1\left[k_1e_1 + \dot{\omega}_{ref} - \mu\lambda_{2d}i_{1q} + \frac{T_L}{J}\right] \\ &\quad + e_3\left[k_3e_3 + \left[\dot{\lambda}_{ref} + \alpha R_2\lambda_{2d} - \alpha L_m R_2i_{1d}\right]\right]\end{aligned}\quad (7)$$

Where k_1, k_3 are positive design constants that determine the closed-loop dynamics. Thus the tracking objectives will be satisfied if we choose

$$\begin{aligned}(\lambda_{2d}i_{1q})_{ref} &= \frac{1}{\mu}\left[k_1e_1 + \dot{\omega}_{ref} + \frac{T_L}{J}\right] \\ (R_2i_{1d})_{ref} &= \frac{1}{\alpha L_m}\left[k_3e_3 + \dot{\lambda}_{ref} + \alpha R_2\lambda_{2d}\right]\end{aligned}\quad (8)$$

since the choice of (8) will gives

$$\dot{V} = -k_1e_1^2 - k_3e_3^2 \leq 0 \quad (9)$$

The equations (8) indicate what the virtual controls should be that in order to satisfy our control objectives. So they provide references for the next step of the Backstepping design, which is essentially to make the "torque" and i_{1d} behave in the desired manner.

Now we go to next step and try to make the signals $\lambda_{2d}i_{1q}$ and i_{1d} behave as desired. So we define again error signals involving the desired variables in (8)

$$\begin{aligned}e_2 &= (\lambda_{2d}i_{1q})_{ref} - \lambda_{2d}i_{1q} \\ &= \frac{1}{\mu}\left[k_1e_1 + \dot{\omega}_{ref}\right] + \frac{T_L}{\mu J} - \lambda_{2d}i_{1q} \\ e_4 &= (R_2i_{1d})_{ref} - R_2i_{1d} \\ &= \frac{1}{\alpha L_m}\left[k_3e_3 + \dot{\lambda}_{ref} + \alpha R_2\lambda_{2d}\right] - R_2i_{1d}\end{aligned}\quad (10)$$

Then the error equations (5) can be expressed as

$$\begin{aligned}\dot{e}_1 &= -k_1e_1 + \mu e_2 \\ \dot{e}_3 &= -k_3e_3 + \alpha L_m e_4\end{aligned}\quad (11)$$

Also, the dynamical equations for the error signals e_2, e_4 can be computed as

$$\begin{aligned}\dot{e}_2 &= \phi_1 - \frac{1}{\sigma L_1}\lambda_{2d}u_{1q} \\ \dot{e}_4 &= \phi_2 - \frac{1}{\sigma L_1}R_2u_{1d}\end{aligned}\quad (12)$$

Where $\phi_i' (i=1,2)$ are known signals that can be used in the control, and their expressions are as follows

$$\begin{aligned}\phi_1 &= \frac{k_1}{\mu}(-k_1e_1 + \mu e_2) + \frac{1}{\mu}\ddot{\omega}_{ref} + \eta_1\lambda_{2d}i_{1q} \\ &\quad + \beta n_p\omega\lambda_{2d}^2 + n_p\omega\lambda_{2d}i_{1d} + R_2(\alpha + \eta_2)i_{1q}\lambda_{2d} \\ \phi_2 &= \frac{k_3}{\alpha L_m}(-k_3e_3 + \alpha L_m e_4) + \frac{1}{\alpha L_m}\ddot{\lambda}_{ref} - \frac{\alpha R_2^2}{L_m}(\lambda_{2d} - L_m i_{1d}) \\ &\quad - R_2(-\eta_1i_{1d} + n_p\omega i_{1q}) - R_2^2\left(-\eta_2i_{1d} + \alpha\beta\lambda_{2d} + \alpha L_m \frac{i_{1q}^2}{\lambda_{2d}}\right)\end{aligned}\quad (13)$$

Now we extend the Lyapunov function in (6) to include the state-variables e_2, e_4 as

$$V_e = \frac{1}{2}[e_1^2 + e_2^2 + e_3^2 + e_4^2] \quad (14)$$

We use V_e to derive both the control algorithm. To this end, we again compute the derivate of V_e along with the error equations (11), (12)

$$\begin{aligned}\dot{V}_e &= e_1\dot{e}_1 + e_2\dot{e}_2 + e_3\dot{e}_3 + e_4\dot{e}_4 \\ &= e_1(-k_1e_1 + \mu e_2) \\ &\quad + e_2\left(\phi_1 - \frac{1}{\sigma L_1}\lambda_{2d}u_{1q}\right) \\ &\quad + e_3(-k_3e_3 + \alpha L_m e_4) \\ &\quad + e_4\left(\phi_2 - \frac{1}{\sigma L_1}R_2u_{1d}\right) \\ &= -k_1e_1^2 - k_2e_2^2 - k_3e_3^2 - k_4e_4^2 \\ &\quad + e_2\left[\mu e_1 + k_2e_2 + \phi_1 - \frac{1}{\sigma L_1}\lambda_{2d}u_{1q}\right] \\ &\quad + e_4\left[\alpha L_m e_3 + k_4e_4 + \phi_2 - \frac{1}{\sigma L_1}R_2u_{1d}\right]\end{aligned}\quad (15)$$

From the above we can obtain the control laws as

$$\begin{aligned}u_{1q} &= \frac{\sigma L_1}{\lambda_{2d}}[\mu e_1 + k_2e_2 + \phi_1] \\ u_{1d} &= \frac{\sigma L_1}{R_2}[\alpha L_m e_3 + k_4e_4 + \phi_2]\end{aligned}\quad (16)$$

Witch leads to

$$\dot{V}_e = -k_1e_1^2 - k_2e_2^2 - k_3e_3^2 - k_4e_4^2 \leq 0 \quad (17)$$

3. Adaptive Rotor Flux Estimator

The rotor flux has been estimated using the simplified IM model (3) obtained after the application of the field orientation principle transformation. Where much simplification is gained by the fact that $\lambda_{2q} = 0$. This allows estimation of the rotor-flux vector in polar form as given by

$$\hat{\lambda}_{2d} = -\alpha R_2\hat{\lambda}_{2d} + \alpha L_m R_2i_{1d} \quad (18)$$

4. MRAS Speed Identification

It is based on the MRAS identification approach [17]. Since the motor voltages and currents are measured in a stationary frame of reference, it is also convenient to express these equations in the stationary frame (a,b) [2]:



$$\begin{bmatrix} \dot{\lambda}_{2a} \\ \dot{\lambda}_{2b} \end{bmatrix} = \frac{L_2}{L_m} \begin{bmatrix} u_{1a} \\ u_{1b} \end{bmatrix} - \begin{bmatrix} (R_1 + \sigma L_1 p) & 0 \\ 0 & (R_1 + \sigma L_1 p) \end{bmatrix} \begin{bmatrix} i_{1a} \\ i_{1b} \end{bmatrix} \quad (19)$$

$$\begin{bmatrix} \dot{\lambda}_{2a} \\ \dot{\lambda}_{2b} \end{bmatrix} = \begin{bmatrix} (-1/T_2) & -\omega \\ \omega & (-1/T_2) \end{bmatrix} \begin{bmatrix} \lambda_{2a} \\ \lambda_{2b} \end{bmatrix} + \frac{L_m}{T_2} \begin{bmatrix} i_{1a} \\ i_{1b} \end{bmatrix} \quad (20)$$

Figure1 illustrates the estimation of the motor speed by means of MRAS techniques. Two independent observers are constructed to estimate the components of the rotor flux vector: one based on (19) and the other based on (20). Since (19) does not involve the rotor speed to be estimated, this observer may be considered as a reference model.

Within this approach an identifier consists of a tuning model (20) depending on an estimate of the unknown rotor speed and a mechanism to adjust the estimate. This adjustment is performed to make the output of the tuning model asymptotically match the output of the reference model.

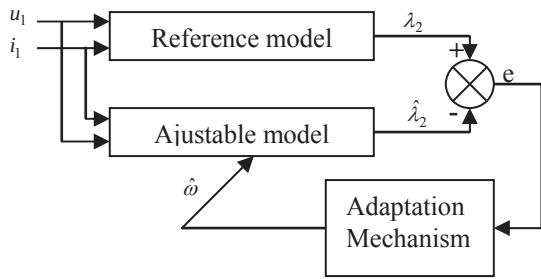


Figure1: Structure of MRAS speed estimator

The adaptive mechanism should be designed to ensure the stability of the control system.

Rotor speed is obtained from the adaptation mechanism as follows:

$$\hat{\omega} = \left(K_p + \frac{K_i}{p} \right) (\lambda_{2b} \hat{\lambda}_{2a} - \lambda_{2a} \hat{\lambda}_{2b}) \quad (21)$$

Figure2 shows the block diagram of the speed identification MRAS based on this adaptation mechanism.

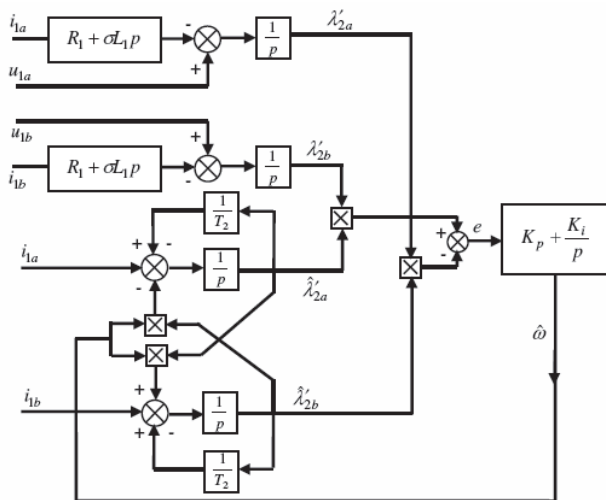


Figure2: Block diagram of MRAS speed estimator

K_p and K_i : are the gain constants of the adaptation mechanism.

5. Stability Analysis

To prove the convergence properties and stability of the overall closed loop system, we will prove that all the signals of the states are bounded and all tracking errors and all estimation and observation errors converge to zero asymptotically. For that, the proof is established for both adaptive Backstepping controller and adaptive MRAS identifier.

a. Adaptive controller:

The proof is established via Barbalat's Lemma using the Lyapunov function candidate [16].

For that, we have to prove

$$e_1, e_2, e_3, e_4 \in \mathcal{L}_\infty \cap \mathcal{L}_2, \quad \dot{e}_1, \dot{e}_2, \dot{e}_3, \dot{e}_4 \in \mathcal{L}_\infty \quad (22)$$

These conditions imply, via Barbalat's Lemma [10] that the closed-loop is stable and that e_1, e_2, e_3 and e_4 all converge to zero.

We begin with the facts that V_e is positive definite, and $\dot{V}_e$ is semi-negative definite, thus all the error signals e_1, e_2, e_3 and e_4 are bounded. With our assumptions on the reference signals $\omega_{ref}, \lambda_{ref}$, we get that the signals ϕ_1 and ϕ_2 are bounded. Therefore by equations (11), (12), we know that $\dot{e}_1, \dot{e}_2, \dot{e}_3$ and $\dot{e}_4$ are bounded. Now, what left is to prove

$$e_1, e_2, e_3, e_4 \in \mathcal{L}_2 \quad (23)$$

By the equation (17) we get

$$\dot{V}_e = -k_1 e_1^2 - k_2 e_2^2 - k_3 e_3^2 - k_4 e_4^2 \leq -k_1 e_1^2 \quad (24)$$

Integrating on both sides from 0 to ∞ we have

$$V_e(\infty) - V_e(0) \leq -\int_0^\infty k_1 e_1^2 dt \quad (25)$$

thus

$$V_e(0) - V_e(\infty) \geq \int_0^\infty k_1 e_1^2 dt \quad (26)$$

which imply that $e_1 \in \mathcal{L}_2$. In the same way, we can also prove that $e_2, e_3, e_4 \in \mathcal{L}_2$.

Therefore, we have concluded that all the signals of the states are bounded and that the error signals e_1, e_2, e_3, e_4 converge to zero asymptotically.

b. Adaptive identifier

For the MRAS identifier design, it is important to take account of the overall stability of the system and to ensure that the estimated quantity will converge to the desired value with suitable dynamic characteristics. Landau [18] has described practical synthesis techniques for MRAS structures based on the concept of hyperstability. When designed according to these rules, the state error equations of the MRAS are guaranteed to be globally asymptotically stable.

Subtracting (20) from (19), we obtain the following state error equations:



$$\frac{d}{dt} \begin{bmatrix} e_a \\ e_b \end{bmatrix} = \begin{bmatrix} (-1/T_2) & -\omega \\ \omega & (-1/T_2) \end{bmatrix} \begin{bmatrix} e_a \\ e_b \end{bmatrix} + \begin{bmatrix} -\hat{\lambda}_{2b} \\ \hat{\lambda}_{2a} \end{bmatrix} (\omega - \hat{\omega}) \quad (27)$$

That is,
$$\frac{d}{dt} [e] = [A][e] - [W] \quad (28)$$

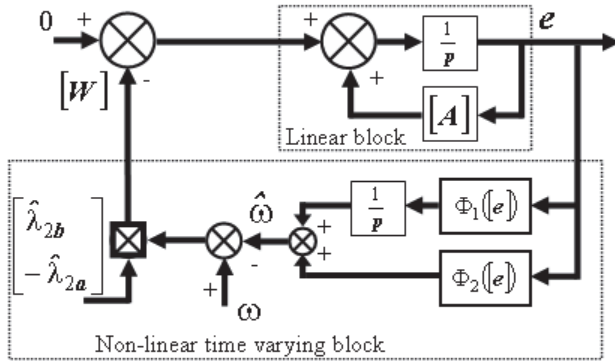


Figure3: Equivalent non-linear feedback system

The estimation algorithm (adaptation mechanism) is chosen according to Popov's hyperstability theory, whereby the transfer function matrix of the linear time invariant subsystem must be strictly positive real and the nonlinear time varying feedback subsystem satisfies Popov's integral inequality, according to which $\int_0^{t_1} (e^T W) dt \geq 0$ in the time interval $[0, t_1]$ for all $t_1 \geq 0$ [19].

Since $\hat{\omega}$, is a function of the state error, these equations describe a nonlinear-feedback system as illustrated in figure3. Following Landau, hyperstability is assured, provided that the linear time-invariant forward-path transfer matrix is strictly positive real and that the nonlinear feedback (which includes the adaptation mechanism) satisfies Popov's criterion for hyperstability. Popov's criterion requires a finite negative limit on the input/output inner product of the nonlinear feedback system. Satisfying this criterion leads to a candidate adaptation mechanism as follows:

Let
$$\hat{\omega} = \Phi_2([e]) + \int_0^t \Phi_1([e]) d\tau \quad (29)$$

Popov's criterion requires that

$$\int_0^{t_1} [e]^T [W] dt \geq -\gamma_0^2 \quad \text{for all } t_1 \geq 0 \quad (30)$$

where γ_0^2 is a positive constant. Substituting for $[e]$ and $[W]$ in this inequality and using the definition of $\hat{\omega}$, Popov's criterion for the present system becomes

$$\int_0^{t_1} \left\{ [e_a \hat{\lambda}_{2b} - e_b \hat{\lambda}_{2a}] \left[\omega - \Phi_2([e]) - \int_0^t \Phi_1([e]) d\tau \right] \right\} dt \geq -\gamma_0^2 \quad (31)$$

Using this expression, it can be shown that Popov's inequality is satisfied by the following functions:

$$\begin{aligned} \Phi_1 &= K_i (e_b \hat{\lambda}_{2a} - e_a \hat{\lambda}_{2b}) = K_i (\lambda_{2b} \hat{\lambda}_{2a} - \lambda_{2a} \hat{\lambda}_{2b}) \\ \Phi_2 &= K_p (e_b \hat{\lambda}_{2a} - e_a \hat{\lambda}_{2b}) = K_p (\lambda_{2b} \hat{\lambda}_{2a} - \lambda_{2a} \hat{\lambda}_{2b}) \end{aligned} \quad (32)$$

6. Simulation Results

The overall configuration of the control system for IM is shown in Figure4. The effectiveness of the proposed controller combined with the rotor speed estimation has been verified by simulations in Matlab/Simulink.

The parameters of the induction motor used are given in Appendix. The simulation results have been obtained under a constant load of 10N.m between (2 and 6)s.

Parameters of the Backstepping controller are:

$k_1 = 28000$, $k_2 = 12000$, $k_3 = 21000$ and $k_4 = 13000$.

Parameters of the MRAS speed identifier are: $K_p = 50000$, $K_i = 100000$.

The simulation results of the performance tracking are demonstrated in the following cases:

Case 1. The speed references are smooth functions, $\omega_{ref} = 50(1 - \exp(-2\pi)) + 10 \sin(2\pi)$ and $50 \sin(2\pi)$

The reference flux is set to 0.5 wb.

As shown in the results of Figure5, the performance are satisfactory.

Case 2. The speed reference is a sort of step type (200 rad/sec) to validate the proposed controller.

The reference flux is set to 0.5 wb.

Figure 6 shows the performance of the proposed scheme with a step speed reference. The errors of speed estimation and speed tracking are rapidly converged.

The control objective is achieved even with the crucial condition.

Case 3. The speed tracking controller is operated in a critical situation of benchmark commands (rapidly changes as 30 – 0 – -30 – 0 – 10 rad/sec).

Figure 7 shows the satisfactory performances of the speed tracking. We can see that the actual speed follows the speed command. Although the speed reference rapidly changes, the actual speed output of the induction motor is closely follows up. On the other hand, the current input is zero as the motor stops.

Thus, the simulation results confirm the robustness of the proposed scheme with respect to the variation of the speed rotor reference.

7. Conclusion

In this paper, a novel scheme for speed and flux control of induction motor using online estimation of the rotor speed has been described. The nonlinear controller presented provides voltage inputs on the basis of only stator currents measurements and guarantees rapid tracking of smooth speed and rotor flux references. MRAS rotor speed estimator has been used. In simulation results, we have shown that the proposed nonlinear adaptive control algorithm achieved desirable tracking performance within a wide range of the operation of the IM. The proposed method also presented robustness properties with respect to the extreme variation of the rotor speed reference. The other interesting features of the proposed method are that it is simple and is easy to implement in real time.

It will be noted that the presence of the pure integrators in MRAS scheme brings the problems of initial conditions and drift. A low pass filter was used to replace the pure integrator [2].



APPENDIX

INDUCTION MOTOR DATA

Stator resistance	1.34 ohms;
Rotor resistance	1.24 ohms;
Mutual inductance	0.17 H;
Rotor inductance	0.18 H;
Stator inductance	0.18 H;
Number of pole pairs	2
Motor load inertia	0.0153 kgm ² ;

8. References

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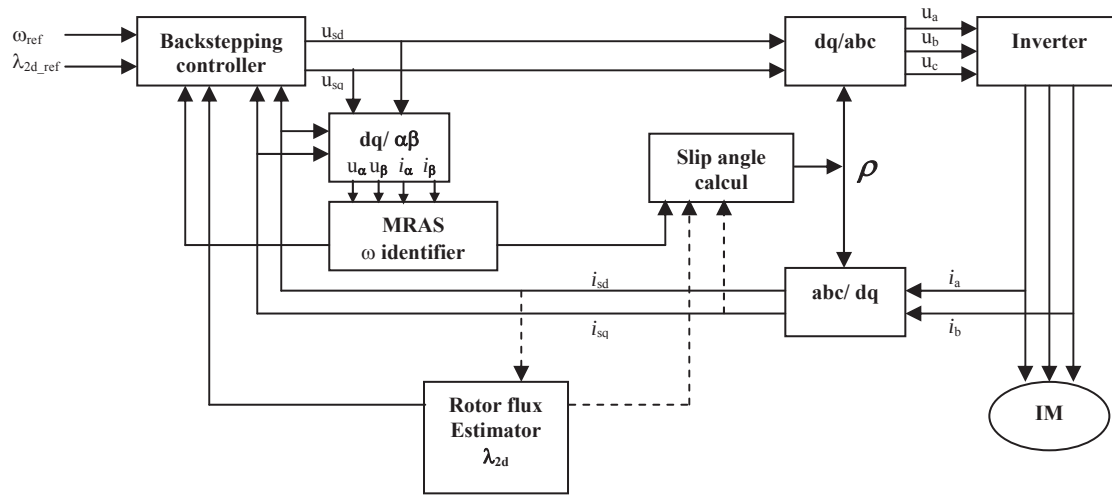


Figure4: Overall bloc diagram of the Sensorless control scheme for IM

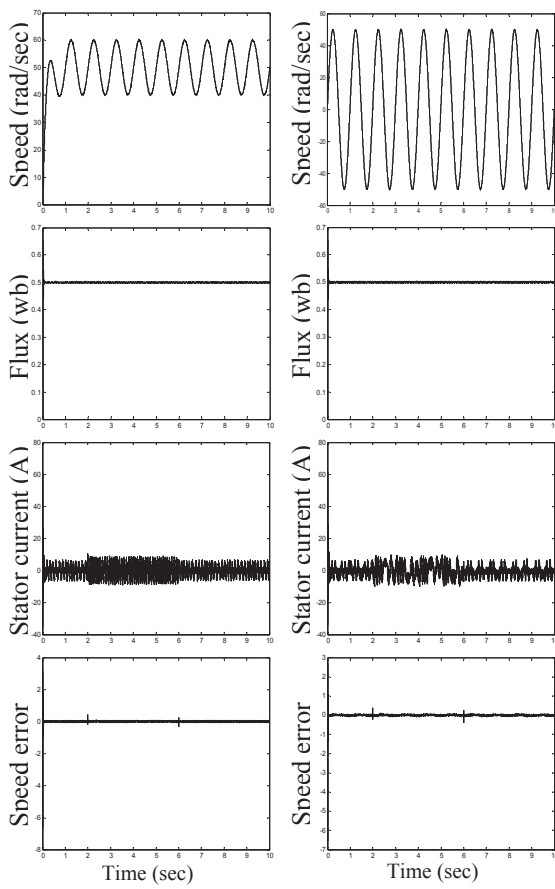


Figure5: Tracking performance

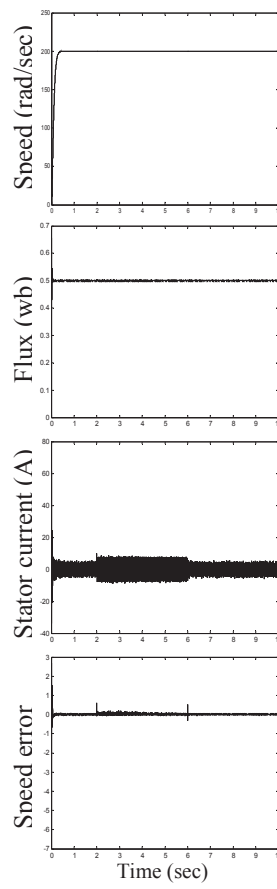
Left : $\omega_{ref} = 50(1 - \exp(-2\pi)) + 10\sin(2\pi)$ Right : $\omega_{ref} = 50\sin(2\pi)$ 

Figure6: Tracking performance

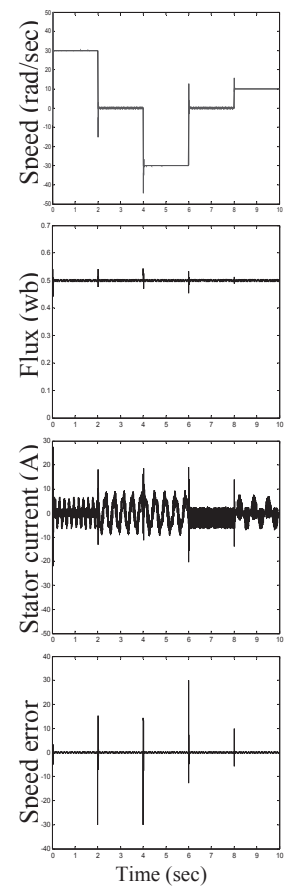
Step type speed tracking
 $\omega_{ref} = 200 \text{ rad/sec}$ 

Figure7: Tracking performance

Benchmark speed tracking
 $\omega_{ref} = (30 \ 0 \ -30 \ 0 \ 10) \text{ rad/sec}$ 

9. Biographies



F. Mehazzem was born in Constantine, Algeria in 1974. He received the engineer degree and the Master degree both in control engineering from the University of Constantine, Algeria His PhD is about the advanced control of electrical systems prepared at ESIEE Groupe-Paris Est University. His research interests includes adaptive and robust nonlinear control and parameters estimation for electrical machines.

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A. Reama was born in Morocco in 1958. He received the Ph.D degree in electrical engineering from National Polytechnic Institute of Toulouse, France, in 1987. His research interests include several aspects of the Electrical Engineering design and control, with particular emphasis on control and optimisation of the real-time distributed systems over an asynchronous communication networks.

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H. Benalla was born in Constantine, Algeria in 1957. He received the B.S., M.S., and Doctorate Engineer degrees in power electronics, from the National Polytechnic Institute of Toulouse, France, respectively in 1981, 1984. In 1995, he received the Ph.D. degree in Electrical Engineering from University of Jussieu-Paris VI, France. His current research field includes

Active Power Filters, PWM Inverters, Electric Machines, and AC Drives.

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A Robust Decentralized Control with Unknown Interconnections

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Abstract

A completely decentralized adaptive fuzzy control procedure is proposed for a class of large-scale nonlinear systems with strong interconnections. The feedback and adaptive controller for each subsystem depend only upon local measurements to provide asymptotic tracking of a reference trajectory. The nonlinearity is approximated using a fuzzy logic method which transforms the nonlinear problems to linear ones. The unknown interconnections and external disturbances are predicted using the orthogonal basis of Hermite. The result of this prediction is taken into consideration in the synthesis of the control law. The proposed controller is combined with a Lyapunov analysis to ensure the stability and convergence. The proposed adaptive algorithm is tested on simulation example to control an interconnected double inverted pendulum.

Keywords: Decentralized Control, Fuzzy Controller, Interconnected Nonlinear System, Orthogonal Basis.

1. Introduction

Decentralized adaptive control systems often arise from various complex situations where there exist physical limitations on information exchange among several systems for which there is insufficient capability to have a single control controller, and due to the physical configuration and high dimensionality of interconnected systems a centralized control is neither economically feasible nor even necessary. Therefore, the decentralized scheme is preferred in control design of large-scale interconnected system [1, 2, 6, 14]. To control a large-scale system, one essential problem is how to handle the interactions among different systems. Intensive research has been devoted to the observer design for large-scale systems. Uncertainties in a large-scale system require the adaptive decentralized technique, for which many decentralized adaptive schemes have been developed, including the model reference adaptive control [1,3], and nonlinear control with a special class of interconnections [5].

These approach focus on stabilization, where the dynamics of subsystems are assumed to be known or to be linear with a set of unknown parameters. However, in

practice, large-scale systems may contain significant uncertainties, and/or with unknown parameters in nonlinear forms and unknown structures.

Fuzzy logic control as one of the most useful approaches for utilizing expert knowledge has been an active field of research the past decade [8, 9, 10]. Fuzzy logic control is generally applicable to plants that are mathematically poorly modeled and where experienced operators are available for providing qualitative guidance. The most important advantage of Fuzzy-logic-control schemes lies in the fact that the developed controllers can deal with increasingly complex systems and controllers without precise knowledge of the model structure of the underlying system dynamic. Recently there have been significant research efforts on these issues in fuzzy control system [8, 10, 11] but these approaches work only for large-scale systems with known or linear dynamics with a set of unknown parameters and interconnection bounds. In practice, however, not all states are usually available.

Based on fuzzy logic control and simple adaptation laws, this paper presents an approach which can easily tackle the output tracking control problem of a class of large-scale nonlinear system with unknown interconnections. The adaptive approach is provided which identifies the isolated subsystem dynamics to produce a stabilizing controller. This approach ensures asymptotic tracking using only local measurement. The interconnection terms and external disturbances are predicted online by using orthogonal basis projection. The stability and convergence are guaranteed using the Lyapunov approach.

The paper is organized as follows: section 2 describes the problem under investigation; in section 3, we introduce the indirect approach and the stability analysis. Simulation results are given to demonstrate the effectiveness of the proposed approach in section 4.

2. Problem Formulation

Consider a class of nonlinear interconnected SISO subsystems S_i ($i = 1, 2, \dots, N$) described as follows:



$$\begin{cases} \dot{x}_{i,1} = x_{i,2} \\ \vdots \\ \dot{x}_{i,n_i} = f_i(x_i) + g_i(x_i)u_i + \Delta_i(\underline{x}) \\ y_i = x_{i,1} \end{cases} \quad (1)$$

where $\underline{x} = [x_1^T, x_2^T, \dots, x_N^T]^T$, $x_i \in \mathbb{R}^{n_i}$ is the global state vector. $u_i(t) \in \mathbb{R}$ and $y_i(t) \in \mathbb{R}$ are respectively the input and output of the subsystem S_i . The functions $f_i(\cdot)$ and $g_i(\cdot)$ are unknown and nonlinear. $\Delta_i(\underline{x}) \in \mathbb{R}$ represent interconnections among unknown subsystems ($i = 1, 2, \dots, N$) and external disturbances.

The tracking error for subsystem S_i is defined by: $e_i = y_{ir} - y_i$. The goal is to design an adaptive control for each subsystem using only states which will cause the output y_i to track a desired output trajectory y_{ir} in presence of strong interconnections and disturbances using only local measurements.

We have to make the following assumptions:

A.1 The vector $\underline{x}_i$ is assumed to be available for measurement, this ensure the observability of each subsystem S_i .

A.2 The controllability is ensured if the functions $g_i(x_i) \neq 0$ ($i = 1, 2, \dots, N$), for each subsystem S_i .

Subsystem (1) can be expressed as:

$$\dot{\varsigma}_i = \Lambda_{i0}\varsigma_i + b_i[f_i(\underline{x}_i) + g_i(\underline{x}_i)u_i + \Delta_i(\underline{x})] \quad (2)$$

where

$$\Lambda_{i0} = \begin{bmatrix} 0 & 1 & 0 & \dots & 0 \\ 0 & 0 & 1 & \dots & 0 \\ \dots & \dots & \ddots & \ddots & \vdots \\ 0 & 0 & 0 & \dots & 1 \\ 0 & 0 & 0 & \dots & 0 \end{bmatrix}, b_i = \begin{bmatrix} 0 \\ \vdots \\ 0 \\ 1 \end{bmatrix} \quad (3)$$

$$\varsigma_i = [\varsigma_{i,1}, \varsigma_{i,2}, \dots, \varsigma_{i,d_i}]^T,$$

$$\varsigma_{i,1} = y_i, \varsigma_{i,2} = \dot{y}_i, \dots, \varsigma_{i,d_i} = y_i^{(d_i-1)}$$

Assume that the given reference y_{ir} is bounded and have up to $d_i - 1$ bounded derivatives. The reference vector is denoted as: $Y_{ir} = [y_{ir}, \dot{y}_{ir}, \ddot{y}_{ir}, \dots, y_{ir}^{(d_i-1)}]^T$. Define the

tracking error of the i^{th} subsystem as: $e_i = y_{ir} - y_i$. Then the error vector of the i^{th} subsystem is given by:

$$\underline{e}_i = [e_i, \dot{e}_i, \dots, e_i^{(d_i-1)}]^T.$$

It is desired that the output error of the i^{th} subsystem follow:

$$e_i^{(d_i)} + k_{i,d_i-1}e_i^{(d_i-1)} + \dots + k_{i,0}e_i = 0$$

here the coefficients are picked so that each:

$$\hat{L} = \varsigma_i^{d_i} + k_{i,d_i-1}\varsigma_i^{d_i-1} + \dots + k_{i,0} \quad (4)$$

has its roots in the open left-half complex plane.

If each subsystem S_i is known and taken independently without disturbances; ($\Delta_i(\underline{x}_1, \underline{x}_2, \dots, \underline{x}_N) = 0$), then the

primary control should be designed to have the following idealized control law [13]:

$$u_i^* = \frac{1}{g_i(\underline{x}_i)}(-f_i(\underline{x}_i) + K_i^T e_i + y_{ir}^{(d_i)}) \quad (5)$$

In practice the function ($f_i(\underline{x}_i)$ and $g_i(\underline{x}_i)$) are unknown, nonlinear and there is any restriction conditions about the nonlinearity. The terms $\Delta_i(\underline{x})$ are unknown and composed of disturbances and interconnections between the subsystems. In this case, the determination of the primary control law is very difficult to implement. To overcome this problem, we propose to estimate the functions ($f_i(\underline{x}_i)$ and $g_i(\underline{x}_i)$) by using the fuzzy logic technique. The unknown term $\Delta_i(\underline{x})$ which regroups the interconnections, high order and disturbances is predicted using the orthogonal basis of Hermit. The result of this prediction is taken into account to synthesise the auxiliary signal.

3. Adaptive fuzzy decentralized control

The process of fuzzy logic approach includes three steps: fuzzification, fuzzy inference and defuzzification. The fuzzification consist to represent the crisp value into grades of membership for linguistic terms of fuzzy sets. The inference problem is to determine the value of the output based on the knowledge base and rule base. The defuzzification consists of converting fuzzified output into crisp value.

Take an universal fuzzy system $\hat{f}_i(\underline{x}_i / \underline{\theta}_i)$ with $\underline{x}_i \in U_{x_i}$ for some compact set U_{x_i} to approximate the uncertain term $f_i(\underline{x}_i)$ where θ_i contains the tunable parameters. Here, the linearly parameterized fuzzy model [8] is employed in the approximation procedure. Then we replace $f_i(\underline{x}_i)$ and $g_i(\underline{x}_i)$ by the approximated functions

$$\hat{f}_i(\underline{x}_i / \underline{\theta}_{i1}) \text{ and } \hat{g}_i(\underline{x}_i / \underline{\theta}_{i2}) \text{ respectively.}$$

The fuzzy system is described by a collection of fuzzy rule if-then as follow:

$$R_{\phi_i}^{(l)} : \text{if } x_i^1 \text{ is } F_{i1}^l \text{ and } \dots \text{if } x_i^{n_i} \text{ is } F_{in_i}^l \text{ then } \phi_i^l \text{ is } C_i^l$$

where F_{ij}^l and C_i^l are fuzzy sets respectively in U_i and V_i .

The approximated functions $\hat{f}_i(\underline{x}_i / \underline{\theta}_{i1})$ and $\hat{g}_i(\underline{x}_i / \underline{\theta}_{i2})$ can be expressed as:

$$\hat{f}_i(\underline{x}_i / \underline{\theta}_{i1}) = \underline{\theta}_{i1}^T \phi_i(\underline{x}_i) \quad (6)$$

$$\hat{g}_i(\underline{x}_i / \underline{\theta}_{i2}) = \underline{\theta}_{i2}^T \phi_i(\underline{x}_i), (i = 1, 2, \dots, N)$$

where $\underline{\theta}_{ik} = [\theta_{ik}^1, \theta_{ik}^2, \dots, \theta_{ik}^{m_i}] \in \mathbb{R}^{m_i}$ is a parameter vector

($k = 1, 2$) and $\phi_i(\underline{x}_i) = [\phi_i^1(\underline{x}_i), \phi_i^2(\underline{x}_i), \dots, \phi_i^{m_i}(\underline{x}_i)] \in \mathbb{R}^{m_i}$ is a regressive vector. We have using the singleton fuzzifier, center average defuzzifier and product

inference to calculate the regressor $\phi_i^l(\underline{x}_i)$:



$$\varphi_i^l(\underline{x}_i) = \frac{\prod_{j=1}^{n_i} \mu_{F_{i,j}^l}(x_i^j)}{\sum_{l=1}^{m_i} \prod_{j=1}^{n_i} \mu_{F_{i,j}^l}(x_i^j)} \quad (7)$$

where $\mu_{F_{i,j}^l}(\cdot)$ is the membership functions for $1 \leq l \leq m_i$ (m_i is the number of rules) and $1 \leq j \leq n_i$.

In this paper, we propose the decentralized adaptive fuzzy controller defined as:

$$u_i = \frac{1}{\hat{g}_i(\underline{x}_i / \theta_{i2})} (-\hat{f}_i(\underline{x}_i / \theta_{i1}) + K_i^T e_i + y_{ir}^{(d_i)} + u_{is} + u_{ih}) \quad (8)$$

u_{is} and u_{ih} are the auxiliary controllers. u_{is} is introduced to perform the main control action and u_{ih} is synthesised in order to compensate the effect of interconnections. According to the universal approximation theorem [4], there exist optimal approximation parameters θ_{i1}^* and θ_{i2}^*

such that $\hat{f}_i(\underline{x}_i / \theta_{i1}^*)$ and $\hat{g}_i(\underline{x}_i / \theta_{i2}^*)$ can respectively approximate $f_i(\underline{x}_i)$ and $g_i(\underline{x}_i)$ as best as possible. Define the optimal parameter vectors and fuzzy approximation error:

$$\begin{aligned} \theta_{i1}^* &= \arg \min_{\theta_{i1} \in \Omega_{i1}} \left\{ \sup_{x_i \in \mathcal{R}^{n_i}} |f_i(\underline{x}_i) - \hat{f}_i(\underline{x}_i / \theta_{i1})| \right\} \\ \theta_{i2}^* &= \arg \min_{\theta_{i2} \in \Omega_{i2}} \left\{ \sup_{x_i \in \mathcal{R}^{n_i}} |g_i(\underline{x}_i) - \hat{g}_i(\underline{x}_i / \theta_{i2})| \right\} \end{aligned} \quad (9)$$

where Ω_{i1} and Ω_{i2} are the convex compact sets which contain parameter sets for θ_{i1}^* and θ_{i2}^* respectively:

$$\begin{aligned} \Omega_{i1} &= \{ \theta_{i1} : \text{tr}(\theta_{i1} \theta_{i1}^T) \leq M_{i1} \} \\ \Omega_{i2} &= \{ \theta_{i2} : \text{tr}(\theta_{i2} \theta_{i2}^T) \leq M_{i2} \} \end{aligned} \quad (10)$$

define

$$\begin{aligned} \Delta f_i(\underline{x}_i) &= f_i(\underline{x}_i) - \hat{f}_i(\underline{x}_i / \theta_{i1}^*) \\ \Delta g_i(\underline{x}_i) &= g_i(\underline{x}_i) - \hat{g}_i(\underline{x}_i / \theta_{i2}^*) \end{aligned} \quad (11)$$

which denote the minimum approximation error.

Substituting (7) into (1), the tracking error dynamic equation can be written as:

$$\begin{aligned} \dot{e}_i &= \Lambda_i e_i + b_i [f_i(\underline{x}_i) - \hat{f}_i(\underline{x}_i / \theta_{i1}) + (g_i(\underline{x}_i) \\ &\quad - \hat{g}_i(\underline{x}_i / \theta_{i2})) u_i - u_{is} - u_{ih} - \Delta_i(\underline{x}_i)] \end{aligned} \quad (12)$$

where

$$\Lambda_i = \begin{bmatrix} 0 & 1 & 0 & \cdots & 0 \\ 0 & 0 & 1 & \cdots & 0 \\ \cdots & \cdots & \ddots & \ddots & \vdots \\ 0 & 0 & 0 & \cdots & 1 \\ -K_{i0} & K_{i1} & \cdots & K_{id_i} & \end{bmatrix}, \quad b_i = \begin{bmatrix} 0 \\ \vdots \\ 0 \\ 1 \end{bmatrix} \quad (13)$$

from (6) and (11), equation (12) can be written as:

$$\begin{aligned} \dot{e}_i &= \Lambda_i e_i + b_i [\Phi_{i1}^T \varphi_i(\underline{x}_i) + \Phi_{i2}^T \varphi_i(\underline{x}_i) u_i \\ &\quad + \Delta f_i(\underline{x}_i) + \Delta g_i(\underline{x}_i) u_i - u_{is} - u_{ih} - \Delta_i(\underline{x}_i)] \end{aligned} \quad (14)$$

where the i^{th} parameter error vector is defined as:

$$\Phi_{i1} = \theta_{i1}^* - \theta_{i1}, \quad \Phi_{i2} = \theta_{i2}^* - \theta_{i2}$$

The problem of the interconnections is relaxed by introducing the orthogonal basis functions [11]. We project the interconnection terms on the Hermite basis functions, $\Delta_i(\underline{x})$ becomes:

$$\Delta_i(\underline{x}) = \Gamma_i^{*T} S_i(t) + w_{\Delta_i} \quad (15)$$

where Γ_i^* is a q_i vector such that $\Gamma_i^* = [\Gamma_{i,1}^* \quad \Gamma_{i,2}^* \quad \cdots \quad \Gamma_{i,q_i}^*]^T$ and w_{Δ_i} is the projection error.

$S_i(t)$ is the Hermite basis of dimension q_i .

Let Γ_i be the estimate of Γ_i^* , u_{is} is chosen as:

$$u_{is} = -\Gamma_i^T S_i(t) \quad (16)$$

From equation (15) and (16), the error dynamics can be represented as:

$$\begin{aligned} \dot{e}_i &= \Lambda_i e_i + b_i [\Phi_{i1}^T \varphi_i(\underline{x}_i) + \Phi_{i2}^T \varphi_i(\underline{x}_i) u_i \\ &\quad + \Delta f_i(\underline{x}_i) + \Delta g_i(\underline{x}_i) u_i - u_{ih} \\ &\quad - \tilde{\Gamma}_i^T S_i(t) - w_{\Delta_i}] \end{aligned} \quad (17)$$

where $\tilde{\Gamma}_i = \Gamma_i^* - \Gamma_i$.

Throughout this section, we need the following assumption:

A.3 There exists a positive function $0 < W_i(\underline{x}_i)$ such that:

$$\int_0^T \|w_i\| \leq W_i(\underline{x}_i) \quad \forall 1 \leq i \leq N \quad (18)$$

with $w_i = \max(|\omega_i|, |w_{\Delta_i}|)$ and $\omega_i = \Delta f_i(\underline{x}_i) + \Delta g_i(\underline{x}_i) u_i$.

The following update laws are defined for the decentralized indirect adaptive controller:

$$\dot{\theta}_{i1} = \eta_{i1} e_i^T p_i b_i \varphi_i(\underline{x}_i) \quad (19)$$

$$\dot{\theta}_{i2} = \eta_{i2} e_i^T p_i b_i \varphi_i(\underline{x}_i) u_i \quad (20)$$

$$u_{ih} = -\frac{1}{2r_i} e_i^T p_i b_i \quad (21)$$

$$\dot{\Gamma}_i = \eta_{\Delta_i} e_i^T p_i b_i S_i(t) \quad (22)$$

$$u_{is} = -\Gamma_i^T S_i(t) \quad (23)$$

where η_{i1}, η_{i2} and η_{Δ_i} are fixed adaptive gains.

Theorem: Consider the nonlinear subsystem (1), if there exists matrices $p_i = p_i^T > 0$ satisfying the Riccati-equations:

$$\Lambda_i^T p_i + p_i \Lambda_i + Q_i - p_i b_i \left(\frac{1}{r_i} - \frac{2}{\rho^2} \right) b_i^T p_i = 0$$

where $Q_i = Q_i^T > 0$, $\rho > 0$ is a prescribed attenuation level, and r_i is a positive parameter to be designed subject to $2r_i \leq \rho^2$, while the control (8) is adapted according to (19)-(23), then the following properties are guaranteed for $i = 1, 2, \dots, N$:

- (i) All the variables of the closed-loop system are bounded
- (ii) The performance tracking is achieved.

Proof: Take the error dynamic equation (14) and consider the Lyapunov function candidate



$$\begin{aligned} \dot{v}_i = & \frac{1}{2} \dot{e}_i^T p_i e_i + \frac{1}{\eta_{i1}} \Phi_{i1}^T \dot{\Phi}_{i1} \\ & + \frac{1}{\eta_{i2}} \Phi_{i2}^T \dot{\Phi}_{i2} + \frac{1}{2\eta_{\Delta_i}} \tilde{\Gamma}_i^T \dot{\tilde{\Gamma}}_i \end{aligned} \quad (24)$$

where each $p_i \in \mathbb{R}_i^{d_i \times d_i}$ is a positive definite and symmetric matrix.

The time derivative of v_i along the error trajectory

$$\begin{aligned} \dot{v}_i = & \frac{1}{2} \dot{e}_i^T p_i e_i + \frac{1}{2} \dot{e}_i^T p_i \dot{e}_i + \frac{2}{\eta_{i1}} \Phi_{i1}^T \dot{\Phi}_{i1} \\ & + \frac{2}{\eta_{i2}} \Phi_{i2}^T \dot{\Phi}_{i2} + \frac{1}{\eta_{\Delta_i}} \tilde{\Gamma}_i^T \dot{\tilde{\Gamma}}_i \end{aligned} \quad (25)$$

Substituting (14) into (25) we obtain:

$$\begin{aligned} \dot{v}_i = & \frac{1}{2} \dot{e}_i^T (\Lambda_i^T p_i + p_i \Lambda_i) e_i + \\ & e_i^T p_i b_i [\Phi_{i1}^T \varphi_i(\underline{x}_i) + \Phi_{i2}^T \varphi_i(\underline{x}_i) u_i \\ & - u_{ih} + \omega_i - w_{\Delta_i}] + \frac{2}{\eta_{i1}} \Phi_{i1}^T \dot{\Phi}_{i1} \\ & + \frac{2}{\eta_{i2}} \Phi_{i2}^T \dot{\Phi}_{i2} + \frac{1}{\eta_{\Delta_i}} \tilde{\Gamma}_i^T \dot{\tilde{\Gamma}}_i \end{aligned} \quad (26)$$

Applying parameters adaptation laws (19)-(23), yields

$$\begin{aligned} \dot{v}_i \leq & -\frac{1}{2} \dot{e}_i^T Q_i e_i - \frac{1}{2} \dot{e}_i^T p_i b_i \left(\frac{2}{\rho^2} - \frac{1}{r_i} \right) b_i^T p_i e_i \\ & + e_i^T p_i b_i w_i - e_i^T p_i b_i u_{ih} \\ \leq & -\frac{1}{2} \dot{e}_i^T Q_i e_i + \frac{1}{2r_i} (e_i^T p_i b_i)^2 \\ & - \frac{1}{\rho^2} (e_i^T p_i b_i)^2 + e_i^T p_i b_i w_i - e_i^T p_i b_i u_{ih} \\ \leq & -\frac{1}{2} \dot{e}_i^T Q_i e_i - \frac{1}{\rho^2} (e_i^T p_i b_i)^2 + e_i^T p_i b_i w_i \\ \leq & -\frac{1}{2} \dot{e}_i^T Q_i e_i - \left(\frac{1}{\rho} e_i^T p_i b_i - \frac{1}{2} \rho w_i \right)^2 \\ & + \frac{1}{4} (\rho w_i)^2 \\ \leq & -\frac{1}{2} \dot{e}_i^T Q_i e_i + \frac{1}{4} \rho^2 w_i^2 \end{aligned} \quad (27)$$

Consider the composite system Lyapunov candidate

$$\begin{aligned} V = & \sum_{i=1}^N \varepsilon_i v_i \text{ where each } \varepsilon_i > 0, \\ \dot{V} \leq & \sum_{i=1}^N \left[\frac{1}{2} \varepsilon_i \dot{e}_i^T Q_i e_i + \frac{1}{4} \varepsilon_i \rho^2 w_i^2 \right] \end{aligned} \quad (28)$$

Denoting:

$$\begin{aligned} W = & \text{diag}(\sqrt{\varepsilon_1} w_1, \dots, \sqrt{\varepsilon_N} w_N), \\ Q = & \text{diag}(\varepsilon_1 Q_1, \dots, \varepsilon_N Q_N) \text{ and } E = (e_1, \dots, e_N) \end{aligned} \quad (29)$$

one has

$$\dot{V} \leq -\frac{1}{2} E^T Q E + \frac{1}{4} \rho^2 W^T W \quad (30)$$

Integrating the above inequality from $t = 0$ to T yields

$$2V(T) + \int_0^T \|E\|^2 dt \leq 2V(0) + \frac{1}{2} \rho^2 \int_0^T \|W\|^2 dt \quad (31)$$

from $V(T)$ and assumptions, it is clear that

$E(t) \in \Omega_E = \left\{ E / E^T Q E \leq 2V(0) + \frac{1}{2} \rho^2 \max_{1 \leq i \leq N} (w_i) \right\}$ So, all the variables are bounded.

4. Simulation results

To show the performance of the presented controller, we consider the double inverted pendulum [7].

The equations which describe the motion of the pendulums are defined by:

$$\begin{aligned} \dot{x}_{11} &= x_{12}, \\ \dot{x}_{12} &= \left(\frac{m_1 g r}{j_1} - \frac{k r^2}{4 j_1} \right) \sin(x_{11}) + \frac{k r}{2 j_1} (l - b) \\ &+ \frac{u_1}{j_1} + \frac{k r^2}{4 j_1} \sin(x_{21}) \end{aligned} \quad (32)$$

$$\begin{aligned} \dot{x}_{21} &= x_{22}, \\ \dot{x}_{22} &= \left(\frac{m_1 g r}{j_2} - \frac{k r^2}{4 j_2} \right) \sin(x_{21}) + \frac{k r}{2 j_2} (l - b) \\ &+ \frac{u_2}{j_2} + \frac{k r^2}{4 j_2} \sin(x_{12}) \end{aligned} \quad (33)$$

where $x_{11} = \phi_1$ and $x_{21} = \phi_2$ are the angular displacements of the pendulums from vertical. The parameters $m_1 = 2kg$ and $m_2 = 2.5kg$ are the pendulum end masses, $j_1 = 0.5kg$ and $j_2 = 0.625kg$ are the moments of inertia, the constant of connecting spring is $k = 100N/m$, the pendulum height is $r = 0.5m$, the natural length of the spring is $l = 0.5m$ and the gravitational acceleration is $g = 9.81m/s^2$.

The distance between the pendulum hinges is defined as $b = 0.4m$ (with $b < l$ in this example, so that the pendulum links repels each other when both are in the upright position. The control parameters for simulation are chosen as follows:

$$\eta_{i1} = 0.001, \eta_{i2} = 0.001, \eta_{\Delta_i} = 0.0001, Q_i = \text{diag}(10, 10),$$

and each Λ_i so that $\hat{L}(\zeta) = \zeta^2 + 4\zeta + 4$ has roots at $(-2, -2)$.

In simulations we are chosen the same conditions as in [12].

$$(x_{11}, x_{12}, x_{21}, x_{22})^T = (\pi/3, 0, -\pi/3, 0)^T, \theta_{i1} = \theta_{i2} = 0_{25 \times 1} \text{ and } \rho = 0.1.$$

We define three fuzzy sets for component of each

$\underline{x}_1 = (x_{11}, x_{12})$ and $\underline{x}_2 = (x_{21}, x_{22})$ with labels $A_{x_{ij}}^1, A_{x_{ij}}^2, A_{x_{ij}}^3, A_{x_{ij}}^4$ and $A_{x_{ij}}^5$ characterized by:

$$\mu_{A_{x_{ij}}^1}(x_{ij}) = \exp(-(x_{ij} + 0.8)^2)$$

$$\mu_{A_{x_{ij}}^2}(x_{ij}) = \exp(-(x_{ij} + 0.4)^2)$$

$$\mu_{A_{x_{ij}}^3}(x_{ij}) = \exp(-(x_{ij})^2)$$



$$\mu_{A_{x_{ij}}^4}(x_{ij}) = \exp(-(x_{ij} - 0.4)^2)$$

$$\mu_{A_{x_{ij}}^5}(x_{ij}) = \exp(-(x_{ij} - 0.8)^2)$$

with $n_{i,j} = 5$, $j = 1, 2$ and $i = 1, 2$.

Defining 25 fuzzy rules, in the following linguistic description:

$$R_i^{(l)} : \text{if } x_{i1} \text{ is } A_{x_{i1}}^{j_1} \text{ and } x_{i2} \text{ is } A_{x_{i2}}^{j_2} \text{ then } y_i^l \text{ is } C_i^l \quad (34)$$

$$\text{Denoting } D_i = \sum_{l=1}^{25} \prod_{k=1}^2 \mu_{A_{x_{ik}}^l}(x_{ik}),$$

$$\varphi_i(x_i) = [\varphi_{i,1}(x_i)/D_i, \varphi_{i,2}(x_i)/D_i, \dots, \varphi_{i,25}(x_i)/D_i]^T$$

In order to test the proposed approach, different simulations are studied. For each, the reference signal is taken equal to zero.

For the first and second simulation the system is not affected by a disturbances and the functions $f_i(x_i)$ and $g_i(x_i)$ are unknown nonlinear. $f_i(x_i)$ and $g_i(x_i)$ are approximated by a fuzzy logic approach. We note that for the first simulation (figure.1) the interconnections are supposed know.

In the second simulation (figure.2) the unknown interconnections are predicted using the orthogonal Hermite basis truncated after three terms.

It can be seen from figure. 1 and figure. 2 that a good tracking is obtained which prove the effectiveness of the predictions.

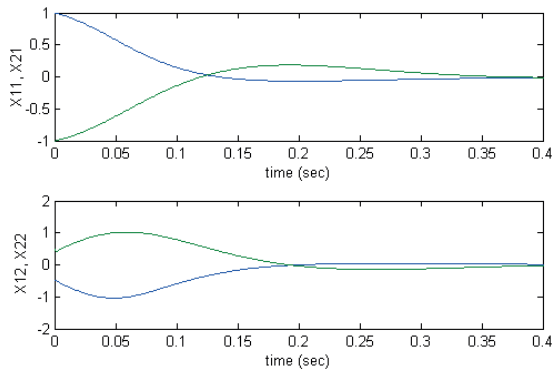


Figure 1. The behaviour of the states: the interconnections are know

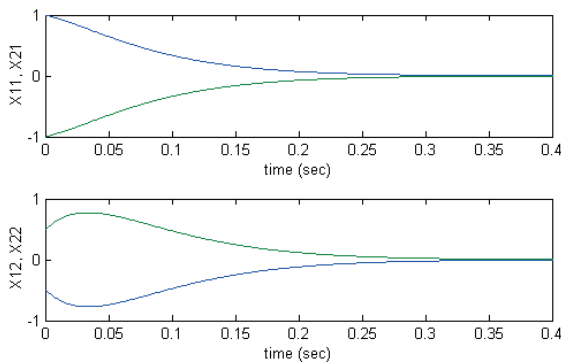


Figure 2. The behaviour of the states: the interconnections are predicted

The third and fourth simulations are realised under the same conditions as for the first and second simulations respectively. A Gaussian noise (figure.5) is injected to the out puts of each subsystem.

For the third simulation, the Gaussian noise is not taken into consideration, figure.3 show that the convergence is less good. When the Gaussian noise is added to unknown interconnections and predicted using the orthogonal basis; a good tracking is illustrated on figure.4.

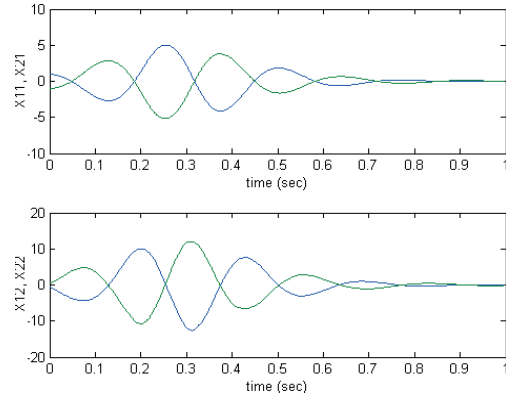


Figure 3. The behaviour of the states: the system is perturbed and the interconnections are know

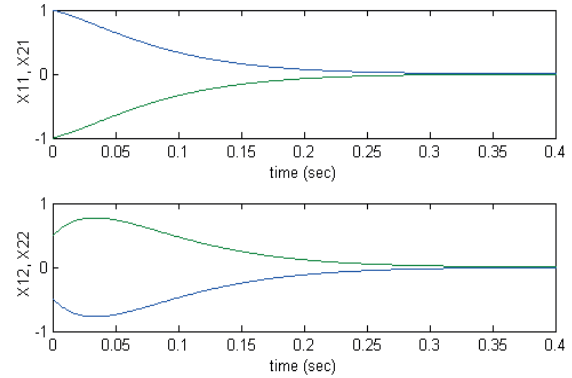


Figure 4. The behaviour of the states: the system is perturbed and the interconnections are unknown

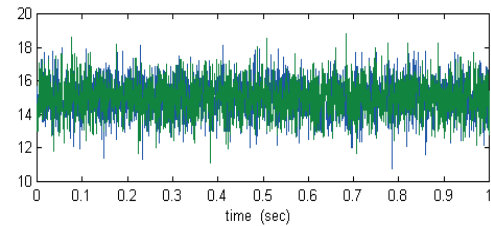


Figure 5. Gaussian noise

5. Conclusion

In this paper, we have presented an adaptive output-feedback fuzzy decentralized control for a class of large-scale nonlinear systems. In the proposed method, fuzzy logic technique is used to estimate the part of the system. The interconnection terms and external disturbances are predicted by projecting them onto orthogonal basis. The adaptive control laws are derived from lyapunov analysis. A good convergence and tracking are obtained without any restrictive conditions on the interconnection terms and disturbances.



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Load Frequency Control with Fuzzy Logic Controller Considering Non-Linearities and Boiler Dynamics

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Abstract

This paper describes the Load Frequency Control (LFC) of two area interconnected reheat thermal system using conventional Proportional – Integral (PI) controller and Fuzzy Logic Controller (FLC). The system is incorporated with governor dead band, generation rate constraint non-linearities and boiler dynamics. The conventional PI controller does not yield adequate control performance with the consideration of non-linearities and boiler dynamics. To overcome this drawback Fuzzy Logic Controller has been employed in the system. The aim of FLC is to restore the frequency and tie line power in very smooth way to its nominal value in the shortest possible time. Time domain simulations are used to study the performance of the power system. System performance is examined considering 1% step load perturbation in either area of the system.

Keywords: Area Control Error, Fuzzy Logic Controller, Governor Dead Band, Load Frequency Control, Proportional-Integral.

1. Introduction

With an increasing demand for electric power, the electric power system becomes more and more complicated. Therefore the supply of electric power with stability and high reliability is required. Normally, the power system operates in normal state which is characterized by constant frequency and voltage profile with certain system reliability. For a successful operation of power system under abnormal conditions, mismatches have to be corrected via supplementary control [4]. Automatic Generation Control (AGC) or Load Frequency Control is a very important issue in power system operation and control for supplying sufficient and reliable electric power with good quality. An interconnected power system can be considered as being divided into control areas, which are connected by the tie lines. In each control area, all generators are assumed to form a coherent group. The power system is subjected to local

variations of random magnitude and duration. For satisfactory operation of a power system the frequency should remain nearly constant. The frequency of a system depends on active power balance. As frequency is a common factor throughout the system, a change in active power demand at one point is reflected through out the system. Mostly the boiler system effects and the governor dead band effects are neglected in the load frequency control studies for simplicity. But for the realistic analysis of system performance, these should be incorporated as they have considerable effects on the amplitude and settling time of oscillations [9], [11].

Due to unnecessary error in the past, conventional PI controller does not provide adequate control performance with the consideration of non-linearities and boiler dynamics. The difficulty in obtaining the optimum settling time of previously said controller is mitigated by using FLC, which gives the opportunity to describe the control action in qualitative term and symbolic form. Literature survey shows that [8], [13-15] only a few investigations have been carried out using FLC to LFC with the consideration of non-linearities and boiler dynamics. It has been discussed that the implementation of FLC has greatly improved the performance of the controller “without negatively affecting the consumers’ quality of supply”. The remainder of the paper is organized as follows: Section (2) focuses on the transfer function modeling of two area thermal system considering boiler dynamics and non-linearities for load frequency control studies. Section (3) emphasizes on conventional proportional-integral (PI) control strategies. Section (4) discusses the proposed fuzzy logic controller. Finally section (5) and section (6) presents the simulation results and conclusion respectively.

2. System Investigated

The detailed block diagram modeling of two area thermal power system, for load frequency control, investigated in this study is shown in figure (1) [5], [6]. Here, area 1 and 2 comprising reheat thermal system with governor dead band, generation rate constraint non-linearities and boiler



dynamics. The nominal parameters of the system are given in Appendix. Matlab version 7.3 has been used to obtain dynamic response such as frequency deviation in area 1 (ΔF_1), area 2 (ΔF_2), tie line power deviation (ΔP_{tie}), and change in steam flow in area 1 (ΔS_{f1}), area 2 (ΔS_{f2}), for 1% step load perturbation in either area of the system.

2.1 Governor Dead Band

Governor Dead Band (GDB) is defined as the total magnitude of a sustained speed change within which there is no resulting change in valve position. The Backlash non-linearity tends to produce a continuous sinusoidal oscillation with a natural period of about 2s [1]. The speed governor dead band has significant effect on the dynamic performance of load frequency control system. Describing function approach is used to incorporate the governor dead band non-linearity [3]. The hysteresis type of non-linearities is expressed as,

$$y = F(x, \dot{x}) \text{ rather than as } y = F(x) \quad (1)$$

To solve the non-linear problem, it is necessary to make the basic assumption that the variable x , appearing in the above equation is sufficiently close to a sinusoidal equation, that is,

$$x \approx A \sin \omega_0 t \quad (2)$$

where, A is amplitude of oscillation

ω_0 is frequency of oscillations

$$\omega_0 = 2\pi f_0 = \pi$$

As the variable function is complex and periodic function of time, it can be developed in a Fourier series as follows, [3]

$$F(x, \dot{x}) = F^0 + N_1 x + \frac{N_2}{\omega_0} \dot{x} + \dots \quad (3)$$

As the backlash nonlinearity is symmetrical about the origin, F^0 is zero. For the analysis in this paper, backlash of approximately 0.05% is chosen [1]. From the above equation, for simplification, neglect higher order terms, the Fourier co-efficients are derived as $N_1=0.8$ & $N_2=-0.2$. By substituting the values in equation (3) the transfer function for GDB is expressed as follows,

$$F(x, \dot{x}) = 0.8x - \frac{0.2}{\pi} \dot{x} \quad (4)$$

2.2 Generation Rate Constraint

In practice, there exists a maximum limit on the rate of change in the generating power. For thermal system a generating rate limitation of 0.1 p.u MW per minute is considered, [10], [13] i.e.

$$\Delta \dot{P}_g \leq 0.1 \text{ p.u. MW / min} = 0.0017 \text{ p.u. MW / s} \quad (5)$$

2.3 Boiler Dynamics

Figure (2) shows the model to represent the boiler dynamics [7]. Boiler is a device meant for producing steam under pressure. The model is basically for a drum type boiler. An oil/gas fired boiler system has been employed in this study, since such boilers respond to load demand changes more quickly than coal-fired units [1], [3]. Drum type boiler is otherwise known as recirculation boiler which relies on natural or forced circulation of drum liquid to absorb energy from the hot furnace walls, called water walls for generating steam.

The boiler receives feed water which has been preheated in the economizer and provides saturated steam outflow. Recirculation boiler make use of a drum to separate steam flow from the recirculation water so that it can proceed to the super heater as a heatable vapour; hence recirculation boiler are referred to as drum type boiler[2]. The changes in generations are initiated by turbine control valves and the boiler controls respond

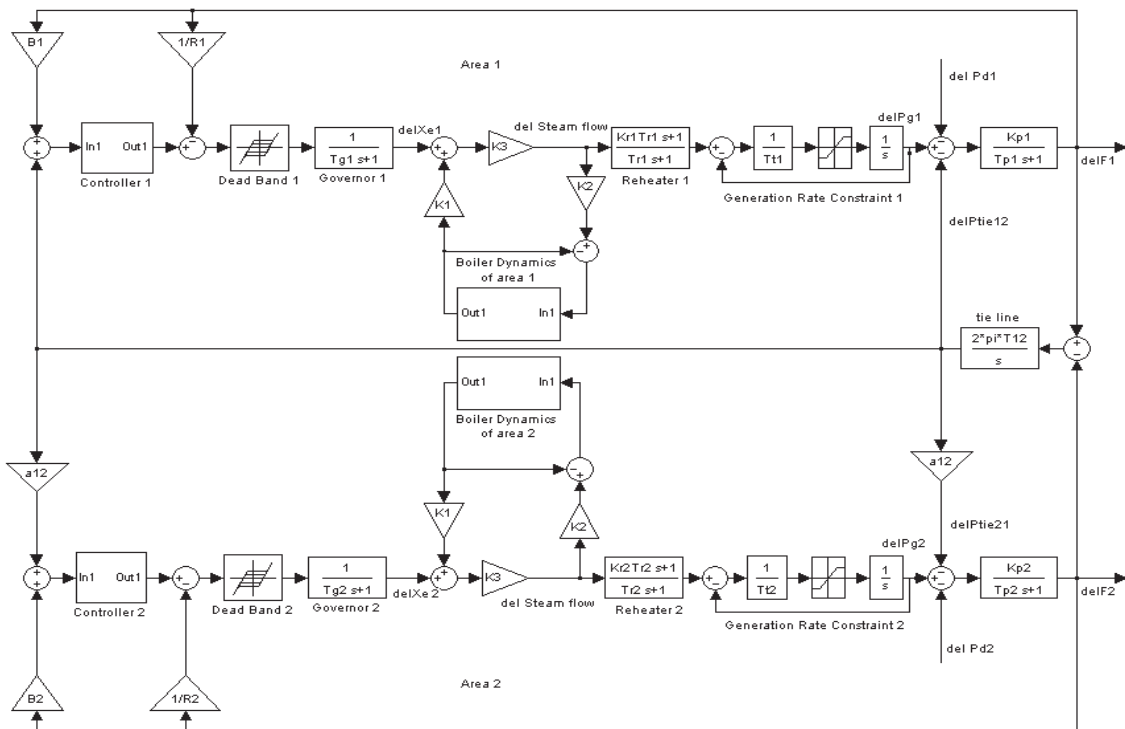


Figure 1: Transfer function modeling of two area thermal system.

with necessary control action, changes in steam flow and changes in throttle pressure, the combustion rate and hence the boiler output.

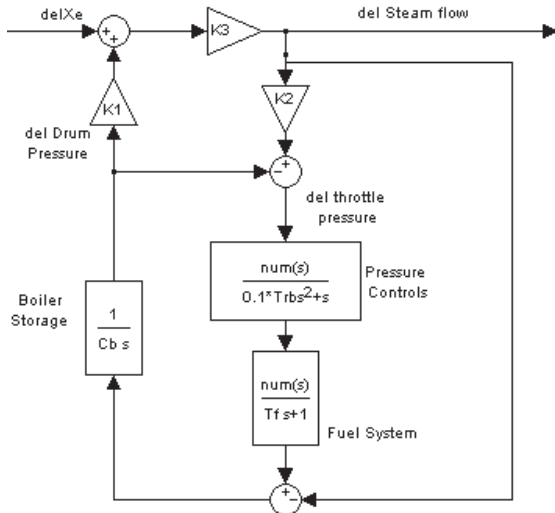


Figure 2: Boiler dynamics

3. Conventional PI Controller

The proportional-integral (PI) controller has received a great deal of attention in the process control areas. It is used as a feedback controller which drives the plant to be controlled with a weighted sum of the error and the integral of that value.

$$u_1 = -K_p \cdot ACE_1 - K_i \int ACE_1 dt \quad (6)$$

$$u_2 = -K_p \cdot ACE_2 - K_i \int ACE_2 dt \quad (7)$$

Where, K_p and K_i are proportional and integral gains respectively, ACE is area control error which defines “a quantity reflecting the deficiency or excess of power within a control area” [17] and u_1, u_2 are controlled output of the respective areas.

The relative simplicity of this controller is a successful approach towards zero steady state error in the frequency of the system. When the integral term is combined with proportional controller it accelerates the movement of the process towards set-point and eliminates the residual steady state error. The conventional control strategy for the LFC problem is to take the error as the control signal. For conventional PI controller the gain K_p and K_i has been determined using Integral Square Error (ISE) criterion. The objective function used for this technique is

$$J = \int_0^t (\Delta F_1^2 + \Delta F_2^2 + \Delta P_{tie}^2) dt \quad (8)$$

where,

ΔF = change in frequency and

ΔP_{tie} = change in tie line power

On the basis of performance index curve as shown in the figure (3) feedback gains should be determined to achieve the optimality of system performance. The main goal of LFC in interconnected power system is to protect the balance between generation and consumption. Because of the complexity and multi variable condition

of the power system, a conventional model may not give satisfactory solution.

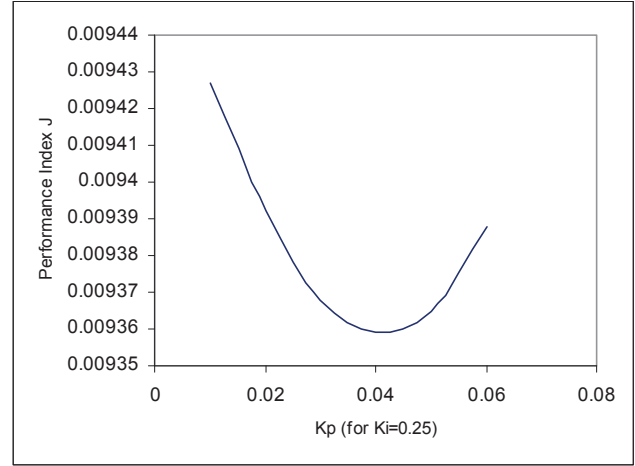


Figure 3: Performance Index Curve

Using the ISE technique for PI controller, the optimum value of K_p and K_i are found to be 0.04 and 0.25 respectively. However, it is learned from the responses with PI controller that, the conventional model responses are unstable. This inference is further reinforced by fuzzy logic controller which is useful in mitigating a wide range of control problems.

4. Fuzzy Logic Controller

The concept of fuzzy logic was developed by Lotfi.A.Zadeh in 1965 to address uncertainty and imprecision which widely exists in engineering problems [12], [15], [16]. His process approach emphasized modeling uncertainties that arise commonly in human thought processes. The design of FLC can be normally divided into three areas namely allocation of area of inputs, determination of rules and defuzzifying of output into a real value. In this paper the proposed fuzzy controller takes the input as ACE and ΔCE , which is given as, [15]

$$ACE_i = \Delta F_i B_i + \Delta P_{tie} \quad (9)$$

The method of fuzzification has found increasing applications in power systems. The applications of fuzzy sets signify a major enhancement of power systems analysis by avoiding heuristic assumptions in practical cases. This is because fuzzy sets could be deployed properly to represent power system uncertainties.

The parameter B_i may be optimized, but here, chosen as, [13]

$$\frac{1}{K_{pi}} + \frac{1}{R_i} \quad (10)$$

The Block diagram of Fuzzy Logic Controller is shown in figure (4). Membership Function (MF) specifies the degree to which a given input belongs to a set. Here, seven membership function have been used to explore best dynamic responses, namely Negative Big (NB), Negative Medium (NM), Negative Small (NS), Zero (ZO), Positive Small (PS), Positive Medium (PM) and Positive Big (PB).

Fuzzy rules are conditional statement that specifies the relationship among fuzzy variables. These rules help us



to describe the control action in quantitative terms and have been obtained by examining the output response to corresponding inputs to the fuzzy controller. Rules are given in Table I. The rules are interpreted as follows,

If ACE is NB and ΔACE is NS then output is PM.

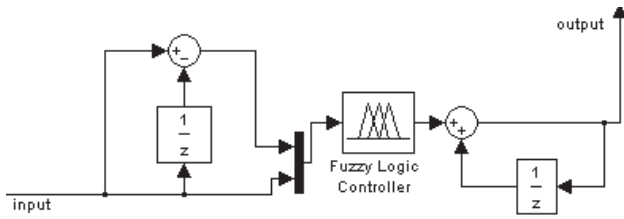


Figure 4: Fuzzy Logic Controller

Defuzzification, to obtain crisp value of FLC output is done by center of area method.

Table I
Fuzzy Control Rules

		ACE							
ΔACE		NB	NM	NS	ZO	PS	PM	PB	
	NB	PB	PB	PB	PB	PM	PM	PS	
	NM	PB	PM	PM	PM	PS	PS	PS	
	NS	PM	PM	PS	PS	PS	PS	ZO	
	ZO	NS	NS	NS	ZO	PS	PS	PS	
	PS	ZO	NS	NS	NS	NS	NM	NM	
	PM	NS	NS	NM	NM	NM	NB	NB	
	PB	NS	NM	NB	NB	NB	NB	NB	

5. Simulation Results

Simulations were performed to a two area reheat thermal system considering non-linearities and boiler effects. The system is simulated for a step load disturbance of 1% on area 2. Due to this, change in dynamic responses of the system has been observed as shown in figure 5(a)-(e). It is observed that the FLC exhibits relatively good performances having very smaller overshoot and transient frequency oscillations. The conventional PI controller does not yield adequate control performance. Simulation result concludes that FLC yields much improved control performance than the conventional PI controller.

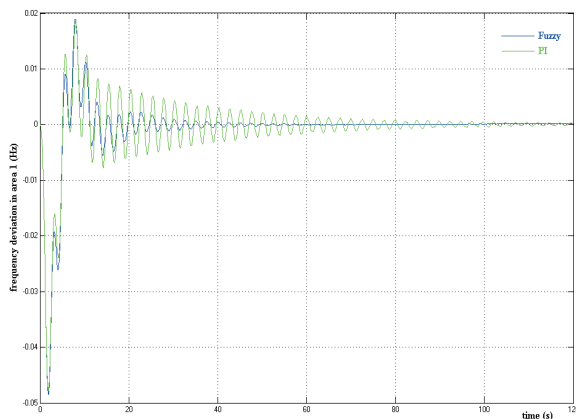


Figure 5(a): Frequency deviation in area 1

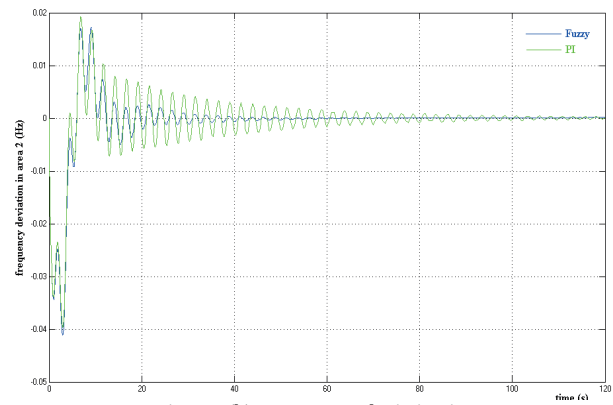


Figure 5(b): Frequency deviation in area 2

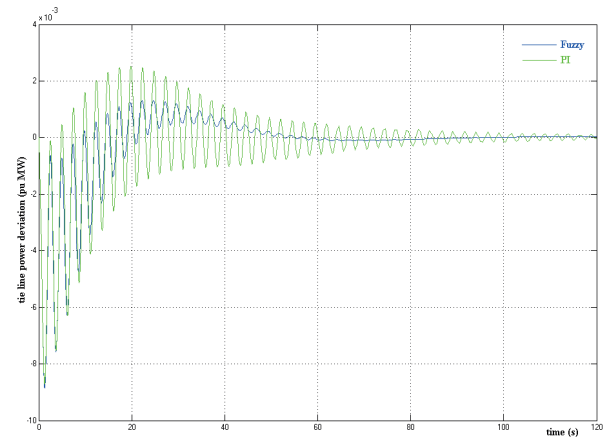


Figure 5(c): Tie line power deviation

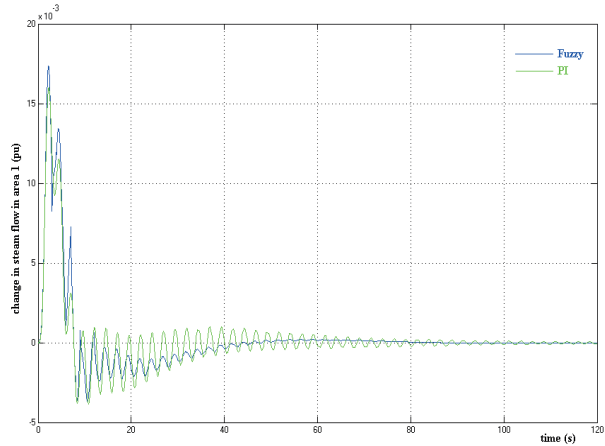


Figure 5(d): Change in steam flow in area 1

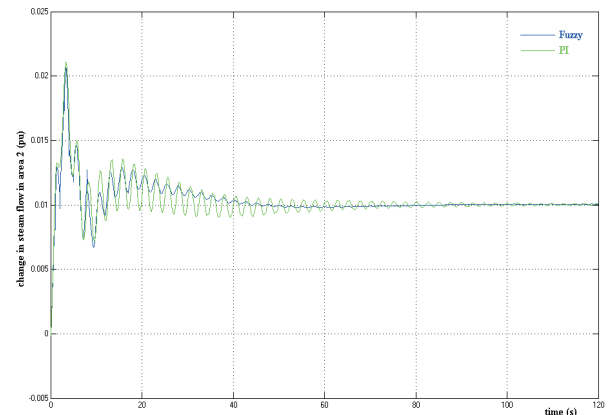


Figure 5(e): Change in steam flow in area 2



6. Conclusion

This study is an application of FLC to load frequency control in the power system. In this work, dynamic behaviour of the frequency of each area, tie-line power deviation and change in steam flow of each area in the power system with two area reheat thermal power system considering non-linearities and boiler dynamics under 1% step load disturbance in area 2, is studied. Integral square error technique has been used to obtain conventional PI controller gains. The conventional PI controller does not yield adequate control performance with the consideration of boiler effects and non linearities. Seven triangular membership function has been used to design FLC. Here, ACE is used as error signal, to control the power system. Finally the simulation results of two-area system with FLC have been compared with conventional PI controller. The simulation results conclude that FLC yields fast settling time with less number of oscillations which advocates the smooth settlement of the quality power supply. Also, the result shows that the FLC yields much improved control performance when compared to conventional PI controller.

7. References

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8. Appendix

(a) System data: [1]

$P_{ti}=2000$ MW
 $T_{ti}=0.3$ s
 $T_{gi}=0.08$ s
 $K_{ti}=0.5$
 $T_{ri}=10$ s
 $K_{pi}=120$ Hz/pu MW
 $T_{pi}=20$ s
 $T_{12}=0.086$
 $R_i=2.4$ Hz/ pu MW
 $f=60$ Hz
 $B_i=0.425$ pu MW/Hz
 $i=1 \text{ \& } 2$



(b) Boiler (oil fired) data: [1]

$K_1 = 0.85$
 $K_2 = 0.095$
 $K_3 = 0.92$
 $C_b = 200$
 $T_d = 0$
 $T_f = 10$
 $K_{ib} = 0.03$
 $T_{ib} = 26$
 $T_{rb} = 69$



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hybrid intelligent system application to power system control problems.





Maximum Power Point Tracking Controller for PV Systems using a PI Regulator with Boost DC/DC Converter

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Abstract

In this paper, we present a novel maximum power point tracking method for a photovoltaic system consisting of a photovoltaic panel with a power electronic converter; the whole is feeding a battery. This maximum power point depends on the temperature and irradiation conditions. A robust control using a PI regulator is used to track this maximum power point. The synthesis of this regulator has been achieved by using Bode method. For having a transfer function of the system, we have used a small signal modeling. Satisfactory theoretical and simulated results are presented. The simulation gives good results.

Keywords: Photovoltaic panel, boost dc/dc converter, PI regulator, Maximum Power Point Tracking.

NOMENCLATURE

MPPT	maximum power point tracking
PV	photovoltaic panel
MPP	maximum power point

1. Introduction

Due to the high cost of solar cells, it is necessary that the PV module operates at its maximum power point. The maximum power produced by a solar cell changes according to the solar radiation and temperature. A PV module is a nonlinear generator. The typical power-voltage characteristic of photovoltaic panel is shown in figure 3. The MPP (V_m , P_m) is reached where $\partial P/\partial V = 0$, $P = VI$ being the PV power. In order to optimize the ratio between the output power and the installation cost, photovoltaic systems have to draw continuously a maximum power from the modules. Maximum power point trackers, commonly known as MPP-trackers, are systems forcing PV modules to operate at their maximum power. Therefore, various solutions are presented in the literature. Many MPPT systems use a microcontroller or a personal computer for implementing sophisticated algorithms [1-4], or even neural networks [5]. These systems ensure very high performances. However they are typically very expensive and often

need a separated, stable power supply for its operation; therefore they are only suitable for high power applications [6]. Another algorithm is based on the searching of operating point which makes $\partial P/\partial V = 0$. Since the sign of $\partial P/\partial V$ gives the direction of the search; it is possible to determine the maximum power operating by continuously detecting the PV power and voltage. In recent years, many MPPT applications based on this searching algorithm have been presented [7]. In [8, 9] an analog controller is proposed, where a boost dc/dc converter is handed for having $\partial I_b/\partial V$ equals zero (Figure 1). In this method, in order to reduce the complexity of the system, we have considered that the battery is equivalent to a constant voltage E on one hand. On the other hand, the converter is assumed to be ideal.

In this paper, we pick up again the work proposed in [9], and we consider that the battery is a constant voltage E in series with a constant resistance R_b [10-12].

A block diagram of the proposed system is shown in figure 1. A boost dc/dc converter, considered to be ideal, is used to interface the PV output to the battery and to track the maximum power point of the PV module. Then the controller must keep $(\partial P/\partial V)$ equals zero. That is possible with action on the duty cycle α ($0 \leq \alpha \leq 1$) according to the solar irradiation λ and to the temperature T. Duty cycle is a signal produced by a PI regulator. For synthesizing this regulator, we have developed a transfer function for the system using the small signal model. Coefficients K_p and K_i of the PI regulator are obtained by frequency synthesis.

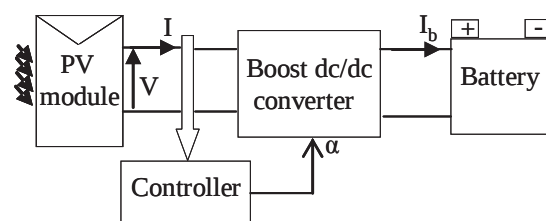


Figure 1: Block diagram of proposed system



The paper is organized as follows: after the Theoretical study and simulation procedure (sections 2 and 3), the theoretical results are presented in section 4 and the simulation results are proposed in section 5. Conclusions are presented in section 6. And finally, the references are given in section 7.

2. Theoretical study

2.1. Optimal operating

The equivalent circuit of a PV module is shown in figure 2. The PV module considered in this paper is the SM55. It has 36 series connected mono-crystalline cells. The relationship between the panel output voltage V and current I is given by authors in [13, 14] as follows:

$$I = I_{sol} - I_{os} \left\{ \exp \left[\frac{q}{\gamma k T} (V + R_s I) \right] - 1 \right\} - \frac{V + R_s I}{R_{sh}} \quad (1)$$

Where:

$$I_{os} = I_{or} \left(\frac{T}{T_r} \right)^3 \exp \left[\frac{q E_{GO}}{\beta k} \left(\frac{1}{T_r} - \frac{1}{T} \right) \right] \quad (2)$$

$$I_{sol} = [I_{sc} + K_I (T - 298.18)] \frac{\lambda}{1000} \quad (3)$$

I , V , I_{os} , T , k , q , K_I , I_{sc} , λ , I_{sol} , E_{GO} , $\gamma (= \beta)$, T_r , I_{or} , R_{sh} , and R_s are respectively, cell output current and voltage, cell reverse saturation current, cell temperature in Kelvin, degree Boltzmann's constant ($1.381e^{-23}$ J/K), electronic charge ($1.602e^{-19}$ C), short-circuit current at 25 °C and 1000 W/m², short-circuit current temperature coefficient at I_{sc} ($K_I = 0.0004$ A/K), solar irradiation in W/m², light-generated current, band gap for silicon ($=1.12$ eV), ideality factor ($=1.740$), reference temperature ($=298.18$ K), cell saturation current at T_r , shunt resistance and series resistance.

The output power of PV panel is $P=VI$, at optimal point, we have:

$$\begin{aligned} \frac{\partial P}{\partial V} = I + V \frac{\partial I}{\partial V} = 0 \Rightarrow \frac{\partial I}{\partial V} = -\frac{I}{V} \\ \Rightarrow I = (V - R_s I) \left\{ I_{os} A \exp[A(V + R_s I)] + \frac{1}{R_{sh}} \right\} \end{aligned} \quad (4)$$

Where: $A = \frac{q}{\gamma k T N_{cell}}$ and N_{cell} is the number of series cells in the module.

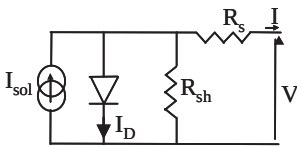


Figure 2: PV panel equivalent circuit

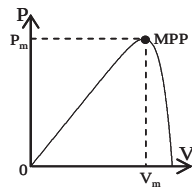


Figure 3: PV panel P-V characteristic

we have measured the shunt resistance R_{sh} and determined the constants in these equations using the manufacturer ratings under standard test conditions of the PV panel.

2.2. Boost converter study

Value L of the inductor required to ensure the converter operating in the continuous conduction mode (figure 5) is

calculated such that the peak inductor current at maximum input power does not exceed the power switch current rating [15]. Hence, L is calculated as:

$$L \geq \frac{V_{om} (1 - \alpha_m) \alpha_m}{f_s |\Delta I_{Lm}|} \quad (5)$$

Where f_s , α_m , ΔI_{Lm} , V_{om} and I_{om} are respectively, switching frequency, duty cycle at maximum converter input power, peak-to-peak ripple of the inductor current, maximum of dc component of the output voltage, and dc component of the output current at maximum output power.

Taking into account that the ripple of the PV output current must be less than 2% of its mean value [15], the input capacitor value is given by:

$$C \geq \frac{I_{om} \alpha_m^2}{0.02 (1 - \alpha_m) V_{inm} f_s} \quad (6)$$

Where V_{inm} is the PV module input voltage at the MPP.

When the boost converter is used in PV applications, the input power, voltage and current change continuously with atmospheric conditions. Thus, the converter conduction mode could change since it depends on them. Also, the duty cycle α is changed continuously in order to track the maximum power point of the PV array. The choice of the converter switching frequency and the inductor value is a compromise between the converter efficiency, the cost, the power capability and the weight.

2.3. Boost converter model

If the chopping frequency is sufficiently higher than the system characteristic frequencies, we can replace the converter with an equivalent continuous model (figure 6). We will consider, for that, the mean values, over the chopping period, of the electric quantities. The transistor had been replaced by a voltage source whose value equals its mean voltage. At the same, the diode had been replaced by a current source. Thus:

$$V_T = E(1 - \alpha) + R_b I_b \text{ and } I_b = (1 - \alpha) I_L \quad (7)$$

Where I_L is the mean value of inductance current.

We deduce from the continuous model equations:

$$C \frac{dV}{dt} = I - I_L \quad (8)$$

$$L \frac{dI_L}{dt} = V - (1 - \alpha)E - R_b I_b \quad (9)$$

At optimal operating point, we have:

$$\begin{cases} I_m = I_{Lm} \\ V_m = (1 - \alpha_m)E + R_b I_{bm} \\ I_{bm} = (1 - \alpha_m) I_{Lm} \end{cases} \quad (10)$$

Where I_m , I_{Lm} , I_{bm} , V_m , α_m are respectively the optimal values of I , I_L , I_b , V and α .

2.4. Small signal model and transfer function

After having put the converter "mean" equivalent circuit in equations, we can determine the transfer function of the system open loop for a small variation around an optimal operating point. Thus, we write $q = q_m + \Delta q$, for each quantity q in the set $\{V, \alpha, I_L, I, I_b, I_{sol}, y\}$



defining the operating point. Where: $y(t) = \frac{\partial P}{\partial V}(t)$. So, the system can shown by a functional diagram like that used in [16] for synthesizing the regulators. The system can be presented as follows:

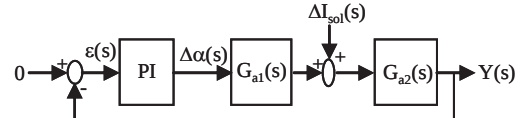


Figure 4: Functional diagram of system.

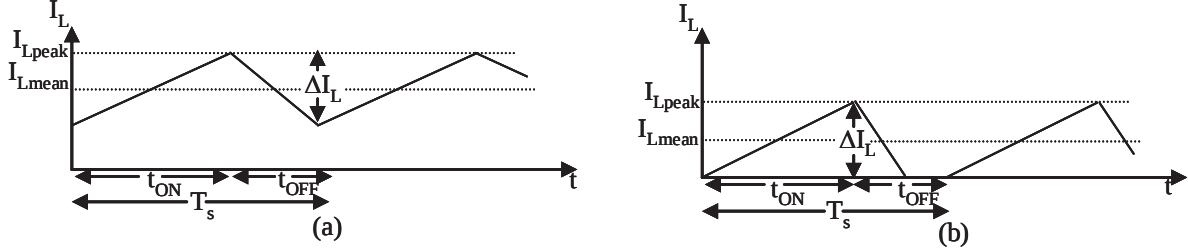


Figure 5: Boost converter waveforms (a) continuous conduction mode (b) discontinuous conduction mode

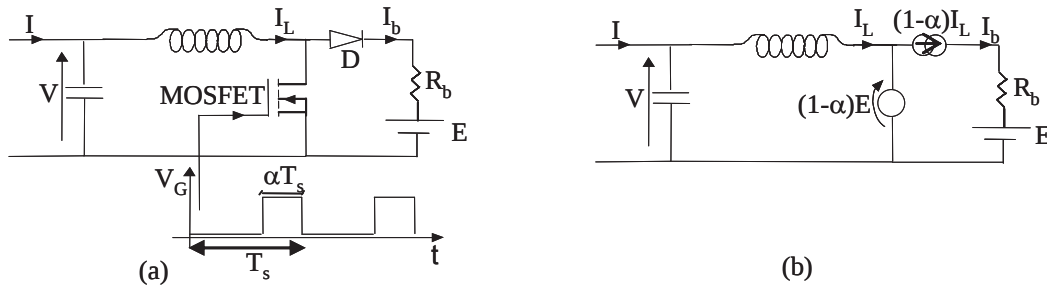


Figure 6: (a) Boost dc/dc converter, (b) boost converter "mean" equivalent circuit

Where:

$$G_{a1}(s) = K_{a1} \frac{1}{b_2 s^2 + b_1 s + 1} \quad (11)$$

$$G_{a2}(s) = K_{a2} \frac{b_2 s^2 + b_1 s + 1}{a_2 s^2 + a_1 s + 1} \quad (12)$$

With:

$$K_{a1} = \frac{(K_1 G - K_2)(E + R_b I_{Lm})(1 + R_s G_m)}{K_1 + K_2 R_b (1 - \alpha_m)}$$

$$K_{a2} = \frac{K_1 + K_2 R_b (1 - \alpha_m)}{(1 + R_s G_m)[1 + G R_b (1 - \alpha_m)]}$$

And:

$$b_2 = \frac{K_1 L C}{K_1 + K_2 R_b (1 - \alpha_m)}; b_1 = \frac{K_1 C R_b (1 - \alpha_m) + K_2 L}{K_1 + K_2 R_b (1 - \alpha_m)}$$

$$a_2 = \frac{L C}{1 + G R_b (1 - \alpha_m)}; a_1 = \frac{C R_b (1 - \alpha_m) + G L}{1 + G R_b (1 - \alpha_m)}$$

$$K_1 = 1 + \frac{A R_s (R_s I_m - V_m)}{R_{dm} (1 + R_s G_m)}; K_2 = \frac{A (R_s I_m - V_m)}{R_{dm} (1 + R_s G_m)} - G$$

$$G = \frac{G_m}{1 + R_s G_m}; G_m = \frac{1}{R_{dm}} + \frac{1}{R_{sh}};$$

$$R_{dm} = \frac{1}{I_{os} A \exp[A(V_m + R_s I_m)]}$$

K_{a1} and K_{a2} are constants gain for a given temperature T and solar irradiation λ .

R_{dm} and G_m are the optimal dynamic resistance and conductance of PV module.

The values of K_{a1} , K_{a2} , K_1 , K_2 , b_2 , b_1 , a_2 , a_1 , G , G_m and R_{dm} can be easily determined if we have know the values of L , C , R_b , E , I_m , V_m , α_m and the parameter values of PV module.

The transfer function of the open loop used to synthesizing a PI regulator is:

$$G_0(s) = G_{a1}(s)G_{a2}(s) = K_{a1}K_{a2} \frac{1}{a_2 s^2 + a_1 s + 1} \quad (13)$$

3. Simulation procedure

We have built our model by using Simulink Matlab. The bloc used for simulations is given by figure 7. In PV module block, equations (1-3) are used; and in block (converter + battery), equations (7-9) are used.

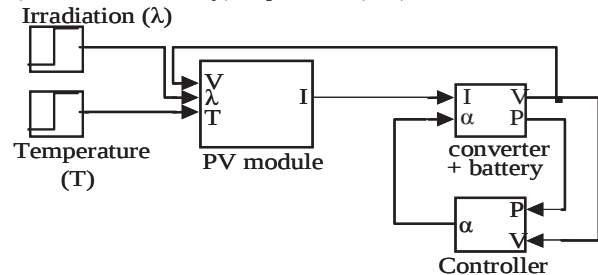


Figure 7: Bloc diagram for system

The proposed controller circuit that forces the system to operate at its optimal operating point under variable



temperature and irradiation conditions is shown in figure 8. Thus, for any temperature and solar irradiation level, the proposed controller circuit is obtained as follows:

- On one hand, we multiply the PV output signal current I by the PV output signal voltage V . We obtain the PV output signal power P who is derived in order to obtain the (dP/dt) signal.
- On the other hand, we derive the signal voltage V , who is inverted. Thus, the signal $1/(dV/dt)$ is obtained.

The product of $1/(dV/dt)$ by (dP/dt) signals, gives the (dP/dV) signal, who is compared to zero. The resulting difference signal (error signal) is used as an input signal of the PI regulator. This PI regulator is used to deliver a duty cycle signal to the dc/dc converter corresponding to the condition: $\frac{dP}{dV} = 0$.

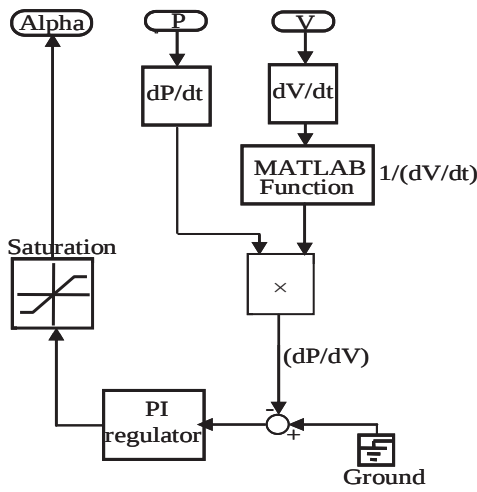


Figure 8: Controller circuit

4. Theoretical Results

4.1. PV module parameters

Measure of R_{sh} . The shunt resistance R_{sh} of PV module (Figure 2) is measured. This measure (Figure 9) is performed in dark room ($I_{sol} = 0$) in laboratory (no wind) by ammeter and voltmeter. We have applied an external negative voltage to the PV panel ($I_D = 0$).

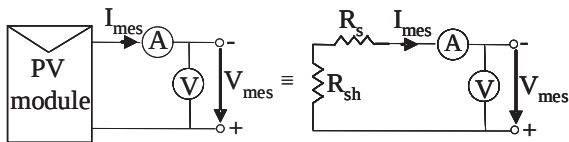


Figure 9: Measure of R_{sh} of PV panel.

The measures of V_{mes} and I_{mes} must be done quickly (no overheating of photocell). The mean value of (V_{mes}/I_{mes}) gives $R_{sh} = 6500\Omega$. R_s is considered negligible in front of R_{sh} .

Determination of the others PV module parameters.

The manufacturer rated values of the SM55 PV module considered in this paper under standard test conditions (irradiation $\lambda = 1kW/m^2$, A.M. = 1.5 solar spectrum and cell temperature $T = 25^\circ C$) are: open-circuit voltage V_{oc}

= 21.7 V, short-circuit current $I_{sc} = 3.45$ A, maximum power current $I_m = 3.15$ A, maximum power voltage $V_m = 17.4$ V and maximum power $P_m = 55$ W.

Using these electrical characteristics, the constants I_{or} , γ ($= \beta$) and R_s in equations (1) and (2), obtained by programming on Matlab Software are $4.842 \mu A$, 1.740 and 0.1124Ω respectively.

4.2. PI regulator and converter parameters.

The PI controller gain and the integral time constant obtained by frequency synthesis using Bode method are respectively $K_p = 0.01$ and $T_i (=1/K_i) = 1.8$ ms. From equation (5), the boost inductance can be chosen as is $L = 1mH$. From equation (6), the input capacitance can be chosen as $C = 4.7\mu F$. The battery voltage and resistance utilized in this paper are respectively $E = 24V$ and $R_b = 0.65\Omega$. Fifty kilohertz switching frequency is adopted.

4.3. Optimal Values at MPP.

For different values of irradiance λ and temperature T , the computation of the theoretical optimum quantities V_m , P_m and α_m of P , V and α are assembled in table 1.

	V_m (V)	P_m (W)	α_m
$\lambda = 100 \text{ W/m}^2$ and $T = 270.18 \text{ K}$	16.77	5.270	0.3069
$\lambda = 1000 \text{ W/m}^2$ and $T = 270.18 \text{ K}$	19.64	62.73	0.2468
$\lambda = 1000 \text{ W/m}^2$ and $T = 320.18 \text{ K}$	15.65	48.61	0.3984
$\lambda = 100 \text{ W/m}^2$ and $T = 320.18 \text{ K}$	12.30	3.714	0.4912

Table 1. Quantities V_m , P_m and α_m for different values of λ and T .

5. Simulation results

In order to demonstrate the effectiveness of the proposed method, some simulations have been carried out.

Simulations were achieved with Matlab Simulink. The simulations studies were made to illustrate the response of the proposed method to rapid temperature and solar irradiance change. For this purpose, the irradiance λ and the temperature T , which are initially $100W/m^2$ and 270.18 K , are switched, at $0.05s$ and $0.15s$, to 1000 W/m^2 and 320.18 K respectively [Figures 10 and 10bis (a), (b) and (c)] and vice versa [Figures 11 and 11bis (a), (b) and (c)], i.e., the solar irradiance changes from 1000 W/m^2 to 100 W/m^2 at $0.05s$ and the temperature changes from 320.18 K to 270.18 K at $0.15s$.

[Figures 10(a), (b) and (c)] and [Figures 11(a), (b) and (c)] give the simulation results for values of K_p and T_i of regulator PI obtained by Matlab programming. After adjust of T_i ($=0.01ms$) we have obtained the [Figures 10bis (a), (b) and (c)] and [Figures 11bis (a), (b) and (c)]. Figures 10(a), 10bis(a), 11(a) and 11bis (a) give the variation of duty cycle, figures 10(b), 10bis(b), 11(b) and 11bis (b) give the variation of the output voltage of PV panel and the figures 10(c), 10bis(c), 11(c) and 11bis(c) give the variation of instantaneous PV power for a step change of temperature and irradiance. It is shown that, the PI controller brings the system into the maximum power point after a some oscillations and the steady state is then reached. At the steady state, it is shown that the average values of voltage, instantaneous power of the PV and duty cycle of a boost DC/DC converter are very



close to their optimal values V_m , P_m and α_m obtained by Matlab programming. So, the relative errors for different average values varied between 0.05% and 0.13%. The simulations of the MPPT show that the system is stable. The oscillations about the computed optimal

operating point are due to the switching action of the DC/DC converter. The transients between operating points are natural for a dynamic system which is controlled by a PI type controller.

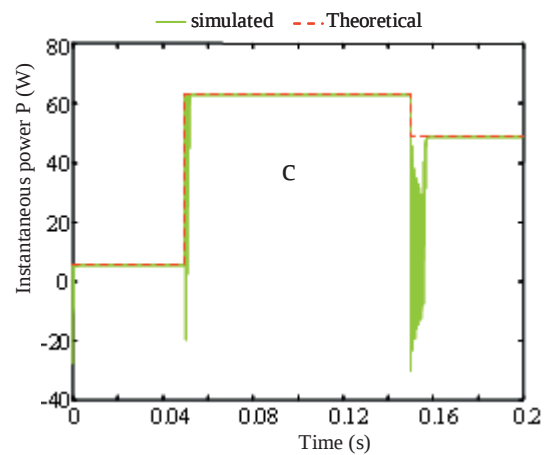
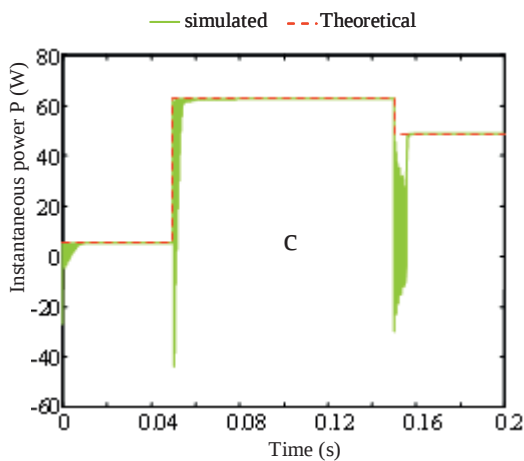
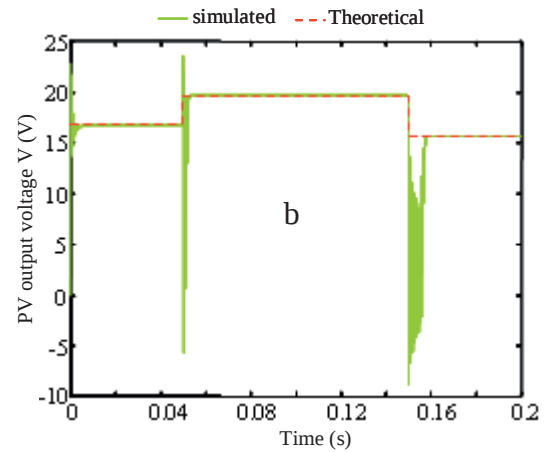
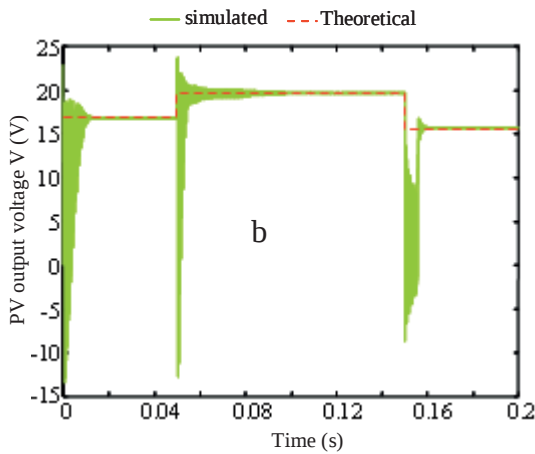
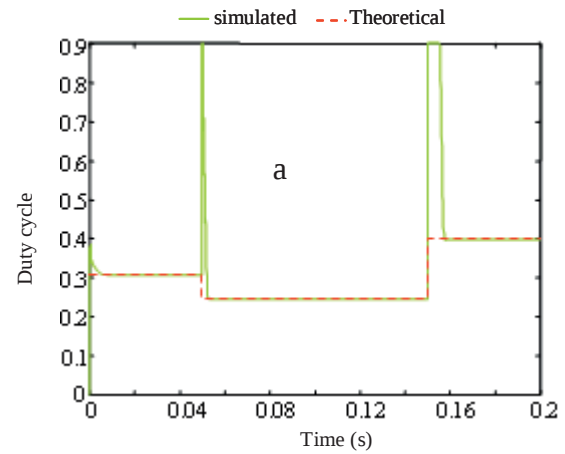
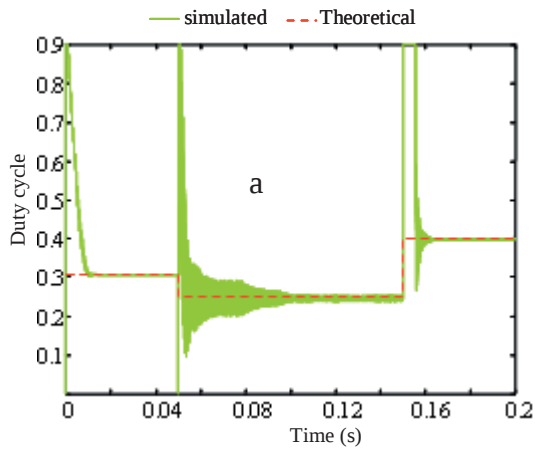


Figure 10: Variation before adjust of T_i , of (a) duty cycle, (b) PV output voltage, and (c) instantaneous PV power for a step change on irradiation λ and temperature T from 100 W/m^2 to 1000 W/m^2 and 270.18 K to 320 K respectively.

Figure 10 bis: Variation after adjust of T_i , of (a) duty cycle, (b) PV output voltage, and (c) instantaneous PV power for a step change on irradiation λ and temperature T from 100 W/m^2 to 1000 W/m^2 and 270.18 K to 320 K respectively.



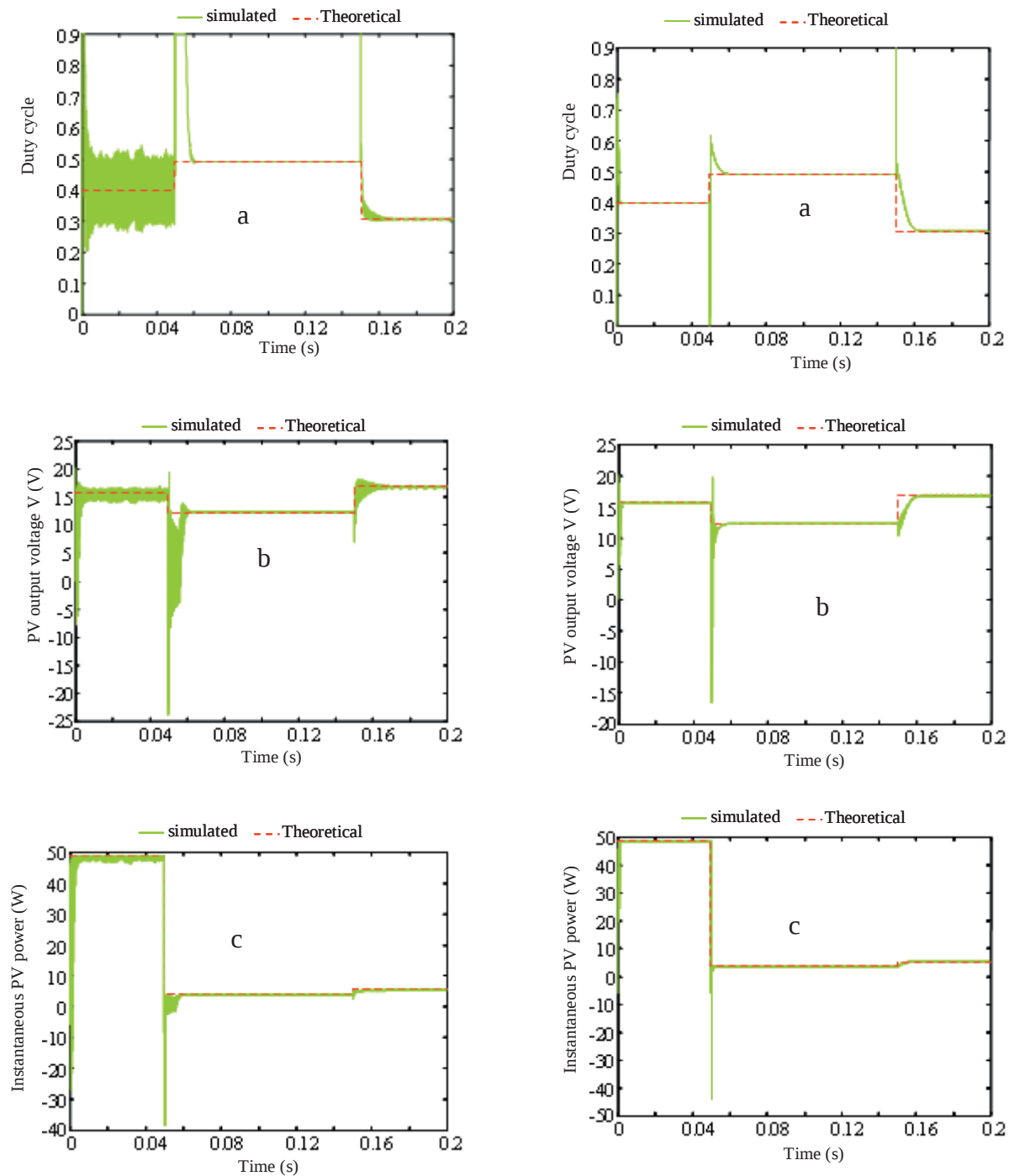


Figure 11: Variation before adjust of T_i , of (a) duty cycle, (b) PV output voltage, and (c) instantaneous PV power for a step change on irradiation λ and temperature T from 1000 W/m^2 to 100 W/m^2 and 320.18 K to 270 K respectively

Figure 11bis: Variation after adjust of T_i , of (a) duty cycle, (b) PV output voltage, and (c) instantaneous PV power for a step change on irradiation λ and temperature T from 1000 W/m^2 to 100 W/m^2 and 320.18 K to 270 K respectively.

6. Conclusion

The output power delivered by a PV module can be maximized using MPPT control system. It consists of a power conditioner to interface the PV output to the load, and a control unit, which drives the power conditioner for extracting the maximum power from a PV array. In this paper, a MPPT system has been proposed and tested by simulations in Matlab Software. It follows the irradiance and the temperature level change rapidly and tracks the MPPT. The PI regulator used to control the boost DC/DC converter is synthesized by frequency synthesis using

Bode method. For that, we have developed a transfer function of global model using a small signal method. Simulations show that the regulation is robust against disturbances.

For a given temperature T and solar irradiance λ , we have calculated the corresponding optimal values α_m , V_m and P_m . The optimal values obtained by programming coincide with ones obtained by simulation. The simulation gives good results.

Our proposed MPPT's charge controller is easier and cheaper to build.



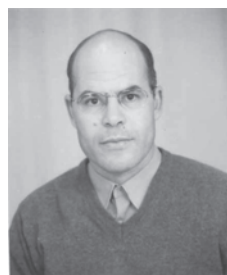
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Acoustic Signal Analysis for Fault Identification of A Control Valve

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Abstract

Acoustic and vibration analysis are well proven non-destructive techniques for the diagnosis of machine health. Control valves are key to quality plant performance. Hence, their life span is of utmost concern for the operators. Condition monitoring for control valves has already been widely researched by several authors using the aerodynamic noise prediction method. This paper describes a method to predict the condition of the control valve, using acoustic signal. Acoustic analysis has been performed by evaluating the dBA levels for determining the faults in control valve. Sub-band approach is used for the acoustic signals to analyze the properties such as amplitude, frequency and energy change. A relationship between the acoustic signature with the faults has been established.

Keywords: *Globe valve, Acoustics, Signal processing, QMF.*

1 Introduction

Control valves play a vital role in the control and optimal performance of a myriad of industrial flow processes. Their precise maintenance of pressures, flow, temperatures, and levels in thousands of installations is a key to overall successful plant performance. However, such stringent requirements and operating conditions induce several faults and failures in control valves. For the past three decades, extensive research is being carried out on predicting failures in control valves. There are many ways for establishing such failure diagnosis systems. Computational Fluid Dynamics (CFD) Modeling has been used to predict the globe control valve performance characteristics. The parameters of interest were flow coefficient and inherent flow characteristics [1, 2]. Erdal and Anderson have used a commercial CFD toolbox to evaluate various numerical effects in

calculating the complex flow through a geometrically simple orifice. In this both the pressure drop and the downstream flow were evaluated and compared with experimental data [3]. From the CFD analysis of steam valve, it was observed that in the partial opening condition a high pressure region revolving circumferentially was the cause of vibrations [4].

Importance of Fault Detection and Diagnosis has been mentioned by A. Asokan and D. Sivakumar in the paper 'Fault Detection and Diagnosis for a Three-tank system using Structured Residual Approach'[5]. The study on Performance Evaluation of Integrated Fault-Tolerant Technique has been carried out by Abulnaja and Saadi[6]. Modeling and Diagnostics of Inductions Machines with neural networks has been proposed by Tarek Aroui et. al.[7].

Online control valve diagnosis has also been widely accepted which uses digital positioners. Kaseda et. al. [8] have developed two algorithms for detecting failures in operating parts of a control valve and one for detecting faulty valve operation due to stick slip while Schruers and Bednar [9] used an expert system to predict faults occurring in the field. For training the experts system, the actual maintenance records of the valve have to be catalogued and stored.

Hyashi et. al. [10] have given detailed mathematical analysis of how the vibrations are generated inside the control valves. With regards to control valve noise, ISA technical committee has drafted a standard for hydrodynamic and aerodynamic noise prediction in control valves. These documents provide the experimental setup and the tools needed to evaluate the noise levels inside the control valve in detail. IEC standard 534-8 7, Control Valve Aerodynamic Noise Prediction, involves first estimating the sound power generated in the fluid inside the valve and piping due to the throttling process [11]. Then, the transmission loss due to the piping is subtracted to determine the sound level at a predetermined



location outside the piping. A computer model has been built to predict the noise in control valve with the help of empirical data collected from experiments at a cogeneration plant [12]. One important issue is the piping vibration generated due to the control valve. Miller looked at the main causes of the flow induced vibration of the piping system. These causes are usually the two-phase flow situations involving cavitation or flashing, standing waves within the pipe, vortex shedding and a high kinetic energy exiting an upstream control valve trim [13]. Piping vibration can be used to measure the aerodynamic noise generated by the control valve. By measuring the vibration generated one meter downstream of a valve, the sound pressure levels can be calculated [14]. Research has been carried out for investigating the noise environments inside butterfly valves [15]. Work also has been done to diagnose the problems related with motorized operated control valves. Sharif et al. have used a pressure sensor along with a valve diagnostic system developed by Fisher Rosemount to perform fault diagnosis. Control valve parameters such as valve travel, frictional force, available seat load and spring rate were also studied [16].

The aim of this work is to predict faults occurring due to mechanical vibrations in control valves, using acoustic approach. For this work, acoustic data is captured using a capacitive microphone. A sound proof enclosure is used to eliminate the noisy process environment. The experimentation is carried on two different set-ups of different line sizes, with two globe valves.

The remainder of the paper is organized as follows: Section (2) focuses on the experiment details for gathering data from the test rigs. The experimental results are provided in the Section (3). Section (4) gives the findings obtained from the experimental results.

2 Experiment Details

To perform diagnosis on control valves, the requirement is to meet the demand for relevant data. The prime aim is to identify the faulty signatures of the control valves from the acoustic data so collected. The important blocks of the experimental set-up as specified in the international standards are given in the Figure 1.

For the acoustic signal capturing, the sensor used is a microphone which converts the sound energy into an electrical output. The output can be directly fed to the sound card of a personal computer and stored for future analysis. A sound proof enclosure is provided around the control valve so as to avoid reflections from the walls and surrounding obstructions. Initially a

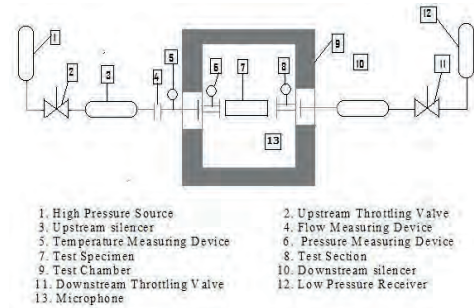


Figure 1: Test Rig as per IEC standard

Sound Level Meter (SLM) is used for validation purpose. IEC standards for control valve noise define the use of anechoic chambers for aerodynamic noise testing of control valves. As interest is more in capturing the noise initiated in the moving parts of a control valve, a sound proof enclosure is used. The microphone is kept as close as possible to the valve to capture the noise generated by the high turbulence region at the vena contracta and the valve trim.

2.1 Microphone

An externally polarized free-field condenser microphone is used to measure the acoustic fields near the control valve. The microphone is placed perpendicular to the control valve axis. Figure 2 shows the locations of the microphone on the control valve. The disturbing effects due to the presence of the microphone in the field are minimal.



Figure 2: Microphone Mounting

2.2 Sound Proof Enclosure

Experiments suggest that due to the diffractions and reflections from the nearby surfaces, the microphone shows higher pressure levels. Standards as discussed earlier require the use of anechoic chamber for measurement of sound pressure level. Since the focus is on mechanical vibrations along the stem and trim assembly, sound proof enclosure is used as shown in Figure



3. This reduces the effects of reflections. The en-

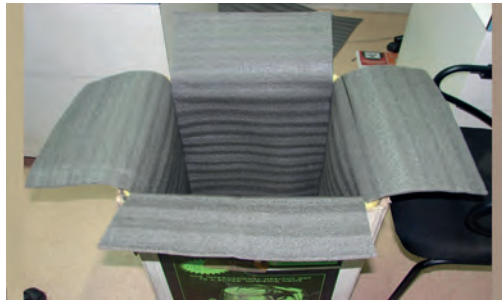


Figure 3: Inner view of the Sound Proof Enclosure

closure is covered with sound proof material from inside to prevent external noise from affecting the signal strength. Sufficient gaps are made in the assembly to allow the piping through the enclosure without affecting the measurement. The details are shown in Figure 4.



Figure 4: Complete Test Assembly

2.3 Test Assembly

The tests are carried out on two different set-ups. The pump specifications for both the set-ups are identical. However, for the first set-up the line size is 25 NB and for the other it is 50 NB.

2.4 Control Valves

Specifications of the globe valves used for the experimentation are listed in Table 1.

Valve 2 is provided with a hand wheel which reduces

Table 1: Valve Specifications

Properties	Valve 1	Valve 2
Valve Size	25 NB	25 NB
Valve Coefficient (Cv)	7.5	4
Actuator	Pneumatic	Hand Wheel
Characteristics	Quick opening	Parabolic

the stem instability observed in the valve 1 due to

pneumatic actuation. Table 2 gives the description of the faults introduced in the valves during experimentation.

Table 2: Control Valve Faults

Valve 1	Valve 2
Increased spring stiffness	Loose Hand Wheel
Decreased spring stiffness	Loose Brass nut
Loose packing	Loose Body nut
Hunting	—

3 Results

The results are presented for the experiments conducted on control valves mounted on two different set-ups.

3.1 Valve 1 on Set-up1

3.1.1 Valve characteristics

Initially the valve characteristics are plotted. The installed valve characteristics for healthy and faulty conditions is as shown in Figure 5.

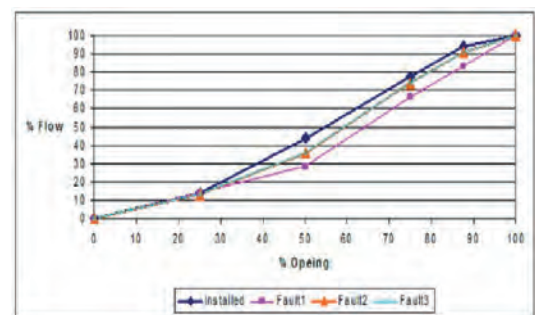


Figure 5: Characteristics of Valve 1

The differential pressure across the Valve with opening of the valve is shown in Figure 6.

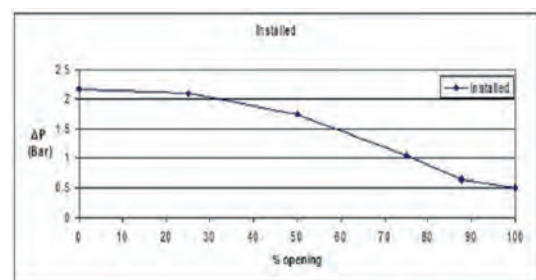


Figure 6: Differential Pressure across the valve 1



3.1.2 Frequency Spectrum

Frequency spectrum plotted from the unprocessed acoustic signal is as shown in Figure 7.

3.1.3 Sound Pressure Level

Table 3 lists the peak dBA values from the recorded signal for valve 1 on the first set-up. The signals are recorded for the healthy as well as for the commonly occurring faults. The nomenclature for various conditions is as follows:

H = Healthy, F1 = Increased stiffness of spring,

F2 = Decreased stiffness of spring,

F3 = Loose packing, F4 = Hunting

Table 3: dBA values for valve 1 on set-up1

Valve opening(%)	H	F1	F2	F3	F4
0	79	83	88	84	105
25	80	86	84	82	98
50	79	85	84	84	—
75	77	84	83	83	—
100	76	84	83	82	—

These faults are true replica of the commonly occurring faults in valves installed in many process industries. There are three types of faults viz. externally correctable, non-correctable or correction needs to be done by dismantling the valve and loop related faults. Representative faults are considered while selecting the faults. Hunting is loop related whereas stiffness change can be corrected externally. For loose packing correction can be done by dismantling the valve.

3.1.4 Energy Distribution

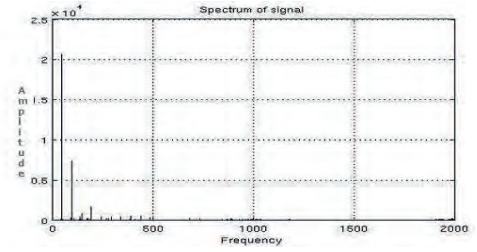
The percentile energy in each frequency band is given in Table 4.

Table 4: % Energy for Valve 1 on set-up 1

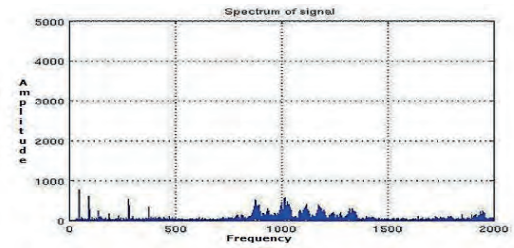
Frequency Band(Hz)	H	F1	F2	F3
215-315	2.75	3	3	5
430-540	0.26	2	1.6	4
860-900	1.27	15	16	9.5
1075-1200	0.1	3	4	3

3.2 Valve 1 on Set-up2

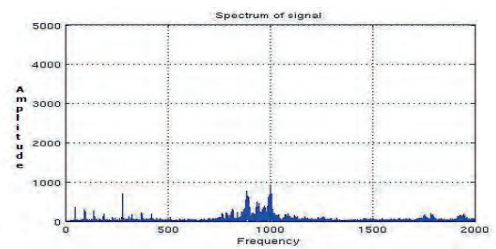
The result of the experiments carried out on valve 1 when installed in a flow loop set-up 2 are as follows.



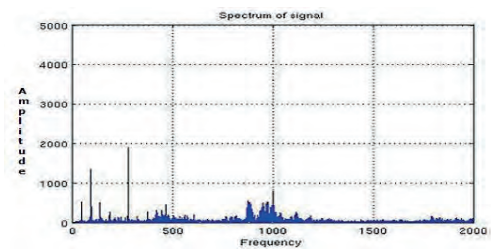
(a) Healthy



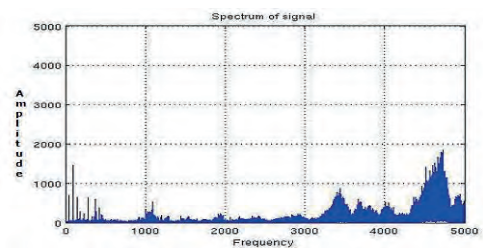
(b) Increased stiffness



(c) Decreased stiffness



(d) Loose packing



(e) Hunting

Figure 7: Frequency spectrum of valve 1 on set-up1



3.2.1 Frequency Spectrum

Spectrum of the recorded acoustic signals is given in Figure 8.

3.2.2 Sound Pressure Level

Table 5 lists the peak dBA values for the valve 1 on set-up2 for healthy as well as faulty conditions.

Table 5: dBA values for valve 1 on Set-up2

Valve opening(in %)	H	F1	F2	F3
0	87	87	85	85
25	86.5	87	84	87
62.5	88.5	87.5	90.5	92
75	91	87.5	89	85.5
87.5	93	90.5	85.5	86
100	87.5	89.5	84	85

3.2.3 Energy Distribution

The percentile energy in each frequency band is given in Table 6.

Table 6: %Energy for Valve 1 on Set-up 2

Frequency Band(Hz)	H	F1	F2	F3
40-45	6.20	2.0	4.50	12.0
85-90	9.50	5.0	7.50	17.5
185-200	8.00	0.3	2.00	1.50
500-525	7.75	0.1	1.75	1.50

3.3 Valve 2 on Set-up1

For the valve 2, the microphone is placed near the stem that is perpendicular to the direction of acoustic source.

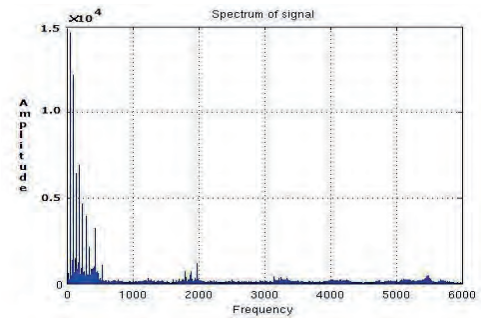
3.3.1 Valve characteristics

The valve characteristics for valve 2 for the healthy and faulty conditions are as show in Figure 9.

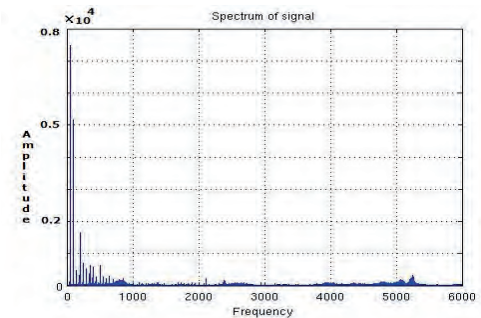
The differential pressure across the Valve with opening of the valve is shown in 10.

3.3.2 Frequency Spectrum

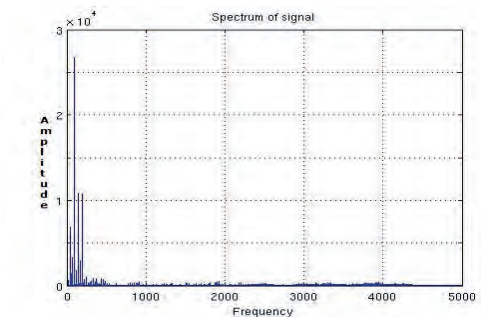
The fourier spectrum of the unprocessed signal and the corresponding faults is shown Figure 11. The nomenclature for various conditions is as follows:



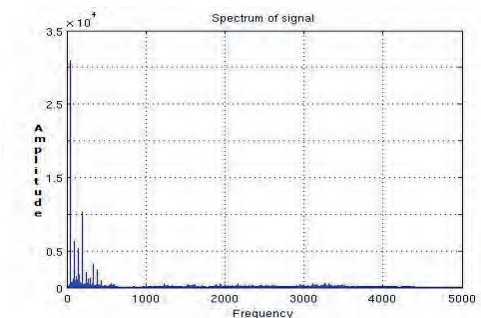
(a) Healthy



(b) Increased stiffness



(c) Decreased stiffness



(d) Loose packing

Figure 8: Frequency spectrum of valve 1 on Set-up2



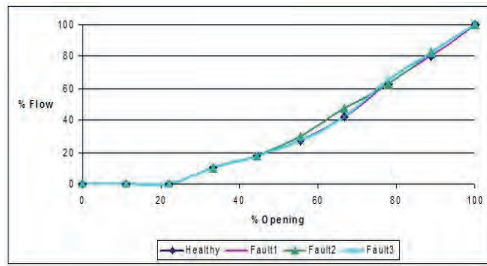


Figure 9: Characteristics of Valve 2

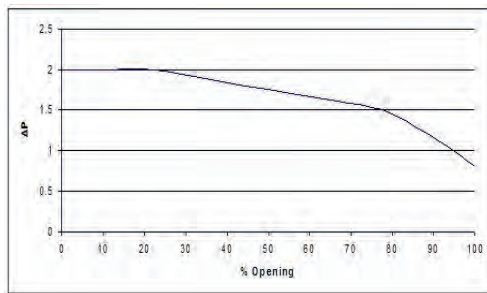


Figure 10: Differential Pressure across the valve 2

H = Healthy, F1 = Loose Hand Wheel,
F2 = Loose Brass Nut, F3 = Loose Body Nut
As the second valve is hand operated, the faults related to valve body are mainly considered. These faults are observed in the environment where structural vibrations are predominant.

3.3.3 Sound Pressure Level

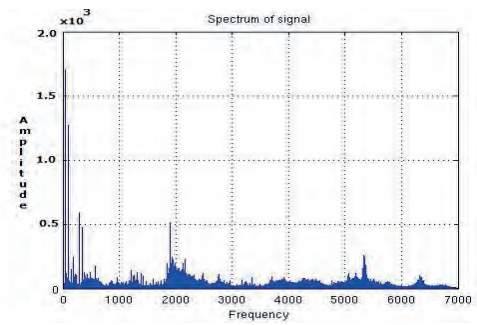
Table 7 gives the details of the dBA values of the valve for healthy and faulty conditions.

Table 7: dBA values of valve 2 on Set-up1

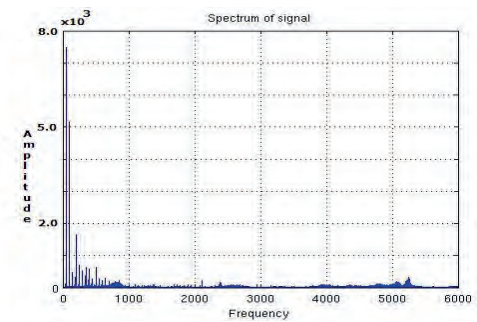
Valve opening(in %)	H	F1	F2	F3
0	69.5	65	64.5	71
22.2	81.5	66	71	70
44.4	89	85	82.5	89
66.6	84	75.5	80	75
88.8	84	85	87	83.5
100	91	85	90.5	90

3.3.4 Energy Distribution

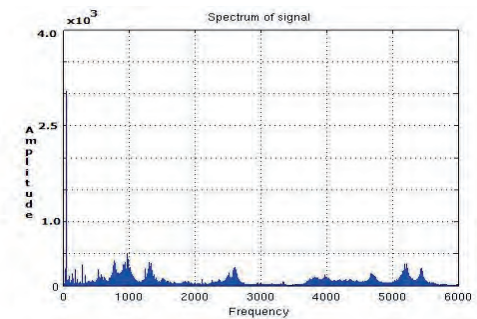
The energy levels in different frequency bands are given in Table 8.



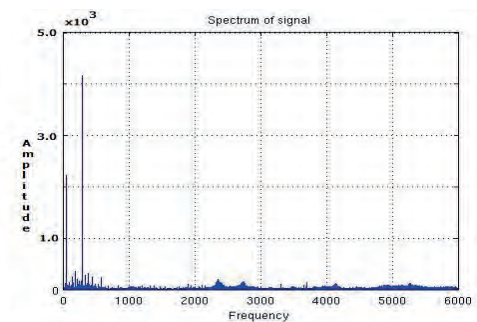
(a) Healthy



(b) Loose Handwheel



(c) Loose Brass Nut



(d) Loose Body Nut

Figure 11: Frequency spectrum of valve 2 on Set-up 1



Table 8: %Energy for Valve 2 on Set-up 1

Frequency Band(Hz)	H	F1	F2	F3
215-430	3.7	4.1	1.6	2.6
1000-1215	3	5.7	11.1	1.3
2150-2365	1.5	3.2	6.63	1.6
3225-3440	2.4	1.25	0.83	2.39

3.4 Valve 2 on Set-up2

The result of the experiments carried out on valve 2 when installed in a flow loop set-up 2 are as follows.

3.4.1 Frequency Spectrum

Spectrum of the recorded acoustic signals is given in Figure 12.

3.4.2 Sound Pressure Level

Table 9 lists the peak dBA values for the valve 2 on set-up 2 for healthy as well as faulty conditions.

Table 9: dBA values for valve 2 on Set-up 2

Valve opening(%)	H	F1	F2	F3
0	60.7	62.75	62.4	65.5
25	62.5	61.75	61.8	68.7
62.5	86	61.8	64.9	70.7
75	65.8	65.5	65.8	71.5
87.5	63.25	65.7	71.3	77
100	79	80.3	78.8	82.9

3.4.3 Energy Distribution

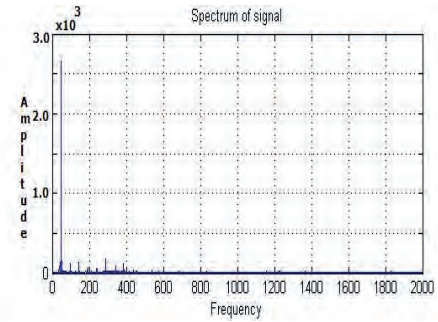
The percentile energy in each frequency band is given in Table 10.

Table 10: %Energy for Valve 2 on Set-up 2

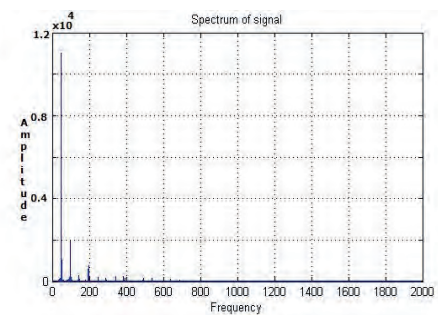
Frequency Band(Hz)	H	F1	F2	F3
40-45	13.5	5.5	4.5	9.5
85-90	0.95	2.0	2.75	0.35
185-200	0.70	1.5	1.25	0.75
500-525	0.5	0.3	0.1	0.75

4 Conclusion

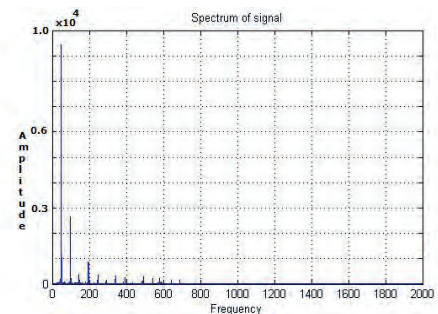
The purpose of this study was to identify the mechanical faults in globe valves using the acoustic approach.



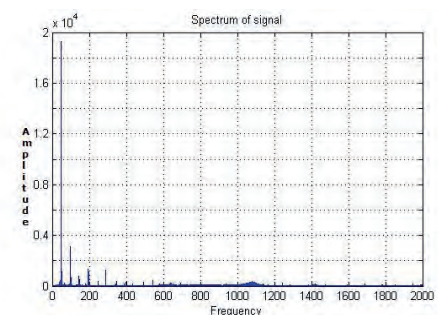
(a) Healthy



(b) Loose Hand Wheel



(c) Loose Brass Nut



(d) Loose Body Nut

Figure 12: Frequency spectrum of valve 2 on Set-up2



Experiments with different aspects for fault identification were carried out. The acoustic data analysis showed a change in the energy and sound pressure levels as regards to healthy and faulty conditions. As observed, each fault generated a pattern different from healthy as well as other faulty conditions. Hence a pattern recognition technique can be used to identify the health status of the valve. The limitation of this technique is the requirement of a sound proof chamber.

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GDPWM Algorithm for Direct Torque Controlled Induction Motor Drive Using the Concept of Imaginary Switching Times

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Abstract

This paper presents a space vector based Generalized Discontinuous Pulsewidth Modulation (GDPWM) algorithm applied to direct torque control (DTC) of induction motor drive without angle estimation using the concept of imaginary switching times. The proposed GDPWM based DTC retains all the advantages of conventional DTC (CDTC) moreover eliminates the complexity involved in conventional space vector Pulsewidth modulation (CSVPWM) algorithm. The proposed method results in reduced current ripple over CSVPWM based DTC at higher modulation indices and low switching losses at all modulation indices. To validate the proposed method, simulation is done in MATLAB/SIMULINK environment and the results of some popular discontinuous PWM (DPWM) based DTC are presented. Also, comparison of the total harmonic distortion (THD) in line current between the CDTC and different DPWM based DTC is made. The proposed algorithm over and above enhancing the performance at high modulation indices, switching losses of the inverter and computational burden involved in conventional GDPWM algorithm are reduced significantly.

Keywords: DTC, CSVPWM, DPWM, GDPWM Imaginary switching times.

1. Nomenclature

R_s, R_r	stator and rotor resistances
L_s, L_r, L_m	stator and rotor self inductances, mutual inductance
P	number of poles
V_{ds}, V_{qs}	d, q axes stator voltages
i_{ds}, i_{qs}	d, q axes stator currents
ψ_{ds}, ψ_{qs}	d, q axes stator flux linkages
ω_r	rotor electrical speed in radians

ω_{sl}	slip speed
ω_e	synchronous speed in electrical radians
T_e	electromagnetic torque
J	inertia constant of the induction motor

2. Introduction

Research interest in high performance control strategies for induction motor drives has grown significantly over the last three decades due to some of their advantages like less maintenance, simple construction and mechanical robustness. Many researchers paid attention to find out different solutions for the induction motor control having the features of precise and quick torque response, and to reduce the complexity involved in FOC [1]. The CDTC has been recognized as viable solution to achieve these requirements [2-6]. There is no need of current controllers in CDTC, as it selects the voltage space vectors according to the errors in flux and torque. It includes flux and torque estimators, flux and torque controllers and a switching table. The switching state of the inverter is updated in every sampling period. With in this sampling period, the inverter does not change its state until the outputs of the controllers change. For this reason the switching frequency is usually not constant; it changes with the rotor speed, load and bandwidth of the two controllers and consequently steady state ripples in torque and flux are high. To solve these problems some research was done on this. Recently several PWM techniques has been developed to reduce the steady state ripple and to get constant switching frequency. One of such techniques is the CSVPWM [7-9]. The steady state performance of the drive with this method is better than the CDTC and also the switching frequency remains fixed at a constant value. Though CSVPWM based DTC could solve most of the problems in CDTC it still has some drawbacks significantly at high modulation indices. Also, the CSVPWM suffers from the drawback of computational burden and increase in memory



requirement. In this paper an attempt is made to avoid the cumbersome calculations thereby eliminating the need of reference voltage vector position and sector calculations using the concept of imaginary switching times [10]-[12]. To improve the performance at high modulation indices, in recent years several DPWM are reported in [14]-[17]. Two popular approaches to PWM generation are the triangular comparison approach and the space vector approach. Performance of a particular DPWM method is modulation index dependent and so an attempt was made in [14],[15] to generalize the developed DPWM methods which lead to the development of generalized DPWM (GDPWM) algorithm based on the triangular comparison method [14].

The main objective of this paper is to propose a space vector based GDPWM algorithm for DTC induction motor drive using the concept of imaginary switching times, which yields high performance drive with reduced complexity. Section 2 presents a survey on the related work. Section 3 gives brief introduction to the CSVPWM algorithm. Section 4 introduces CDTC. Section 5 explains the proposed technique. Section 6 presents the simulation results of different cases of the proposed technique. Finally Section 7 summarizes the present research work stating its merits and demerits.

3. CSVPWM Algorithm

With a 2-level, 3-phase voltage source inverter (VSI) there are eight possible switching states, among which six are active vectors and two zero vectors. It can be shown that all the six active vectors can be represented by space vectors given by (1) forming a regular hexagon and dividing the space plane into six equal sectors denoted as I, II, III, IV, V, VI as shown in Figure1.

$$V_k = \frac{2}{3} V_{dc} * e^{j(k-1)\frac{\pi}{3}}, k = 1, 2, \dots, 6. \quad (1)$$

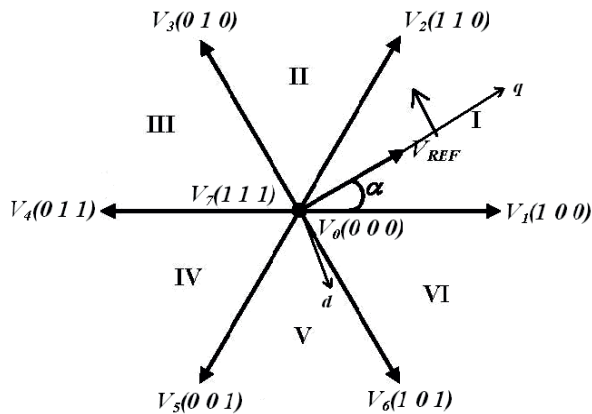


Figure1. Switching states and corresponding voltage vectors of a three phase converter. I, II, III, IV, V, VI are the sectors.

In the CSVPWM strategy, the desired reference voltage vector is generated by time averaging the two nearby active voltage states and two zero states in every sub-cycle T_s . For a given reference voltage V_{REF} at an angle α with reference to V_1 in first sector, the volt-time balance is maintained by applying the active state V_1 , active state V_2 and two zero states V_0 and V_7 together

for durations, T_1, T_2 and T_Z respectively, as given in (2) [3],[10],[14].

$$T_1 = M * \frac{\sin(60^\circ - \alpha)}{\sin 60^\circ} * T_s \quad (2.1)$$

$$T_2 = M * \frac{\sin \alpha}{\sin 60^\circ} * T_s \quad (2.2)$$

$$T_Z = T_s - T_1 - T_2 \quad (2.3)$$

Where 'M' is the modulation index, given by $M = \frac{3V_{REF}}{2V_{dc}}$.

Also in CSVPWM strategy, the total zero voltage vector time, T_Z is divided equally by the two zero vectors V_0 and V_7 . Also in this method, the zero voltage vector time is distributed equally at the start and end of the sampling period in a symmetrical way. Thus, CSVPWM uses $V_0, V_k, V_{k+1}, V_7, V_{k+1}, V_k, V_0$ sequence for odd sectors and $V_0, V_{k+1}, V_k, V_7, V_k, V_{k+1}, V_0$ sequence for even sectors respectively.

4. Principle of CDTC

The principle of CDTC is that rapid instantaneous torque and flux control can be achieved quickly by changing the magnitude and position of stator flux linkage space vector. Decoupled control of torque and flux is achieved by controlling the orthogonal components of stator flux space vector. Neglecting the stator ohmic drop this can be made by properly selecting the inverter voltage states. An optimum switching table is constructed for selecting the appropriate voltage vectors to increase/decrease torque and flux. At every sampling period, based on the error of hysteresis torque and flux controllers an appropriate voltage vector will be selected so as to maintain the torque and flux error within the hysteresis band.

In CDTC since only one vector is applied for the entire sampling period, so for small errors the motor torque may exceed the torque limit. Also because of the use of hysteresis controllers significant ripple in torque and flux is observed in steady state due to bang-bang behavior.

5. Proposed GDPWM Based DTC

The proposed space vector based GDPWM algorithm based DTC induction motor drive combines the principles of CSVPWM algorithm, concept of imaginary switching times, GDPWM algorithm and DTC.

Concept of Imaginary Switching Times: The concept of applying imaginary switching times greatly reduces the complexity involved in CSVPWM algorithm in terms of reference vector position and sector calculation and memory required. With this approach the procedure of identifying the position of the reference voltage vector and sector estimation are eliminated. Assuming that the reference voltage vector is lying in first sector, according to the CSVPWM algorithm it can be best constructed in an average sense by the two of its nearest active voltage vectors (V_1, V_2) and two zero voltage vectors (V_0, V_7) in the specified sequence for a specified time T_1, T_2, T_0, T_7 as in (2) respective [4]. It can be observed that actual



switching times T_1, T_2 are α dependent and so to eliminate the reference voltage vector position dependency the concept of imaginary switching times is applied to CSVPWM algorithm [9-13].

The active vector switching times T_1, T_2 are defined in terms of the imaginary switching times as given in (3)

$$T_1 = T_{an} - T_{bn} \quad (3.1)$$

$$T_2 = T_{bn} - T_{cn} \quad (3.2)$$

The imaginary switching times proportional to the instantaneous values of the reference phase voltages are defined as

$$T_{an} = \left(\frac{T_s}{V_{dc}} \right) V_{an} \quad (4.1)$$

$$T_{bn} = \left(\frac{T_s}{V_{dc}} \right) V_{bn} \quad (4.2)$$

$$T_{cn} = \left(\frac{T_s}{V_{dc}} \right) V_{cn} \quad (4.3)$$

Where, T_s is the sampling period and V_{dc} is the dc link voltage. For every sampling time, the maximum and minimum values of imaginary switching times are calculated and the active vector switching times are expressed in terms of the maximum and minimum switching times.

The effective time is the duration in which the reference voltage vector lies in the corresponding active states, and is the difference between the maximum and minimum switching times. The zero voltage vector switching times is calculated from the switching time duration and effective time as,

$$T_Z = T_s - T_{eff} \quad \text{where,}$$

$$T_{eff} = T_1 + T_2$$

After obtaining the switching times for the active states and zero states the reference voltage vector is synthesized in an average sense within a sub cycle.

GDPWM Algorithm: In recent years several DPWM are reported in [14-15] to improve the performance at high modulation indices. In discontinuous PWM methods zero sequence signals are discontinuous. During each sampling period, one of the phases ceases the modulation and the associated phase is clamped to the positive bus or negative bus. Hence, the switching losses of the associated inverter leg are eliminated. The performance of the PWM methods depends upon the modulation index. In the lower modulation range the continuous PWM (CPWM) methods are superior to DPWM methods, while in the higher modulation range the DPWM methods are superior to CPWM methods. However at all the operating modulation indices, CPWM method has higher switching losses than DPWM methods. Utilizing the generalized phase shift in the DPWM methods, generalized discontinuous PWM (GDPWM) is proposed in [14].

The SVPWM algorithm utilizing the freedom of dividing the zero state time within a sampling period generates different DPWM methods. A zero voltage vector distribution variable σ that divides the zero state time between two zero states are given by

$$T_0 = T_Z \sigma \quad (5.1)$$

$$T_7 = T_Z (1 - \sigma) \quad (5.2)$$

Observation from the eight feasible switching states of the two level, 3-phase VSI, pole voltages V_{ao}, V_{bo}, V_{co} and neutral voltages V_{nok} (k is the switching state) for various switching states lead to the generalized expressions for the pole voltages given by (6).

$$V_{ao} = \frac{1}{2} V_{dc} M_{an} \quad (6.1)$$

$$V_{bo} = \frac{1}{2} V_{dc} M_{bn} \quad (6.2)$$

$$V_{co} = \frac{1}{2} V_{dc} M_{cn} \quad (6.3)$$

where M_{an}, M_{bn} and M_{cn} are the carrier based modulation signals comprising of fundamental frequency components, ranging between 1 and -1. The equations for the modulating signals of the top devices from (6) are expressed by the equation (7).

$$M_{in} = \frac{2V_{in}}{V_{dc}} + \frac{2V_{no}}{V_{dc}}, \quad i=a, b, c \quad (7)$$

The neutral voltage V_{no} can be approximated by time averaging over T_s as given in (8).

$$\begin{aligned} V_{no} &= V_{no1}T_1 + V_{no2}T_2 + V_{no0}T_0 + V_{no7}T_7 \\ &= \frac{V_{dc}}{6}(T_2 - T_1) + \frac{V_{dc}}{2}(1 - 2\sigma)T_Z \end{aligned} \quad (8)$$

Expressing the switching times of the active voltage vectors in terms of the instantaneous maximum V_{max} and minimum V_{min} voltages, equation that is valid for all the six vectors, can be generalized as (9). Using (7) and (9) the expressions for the modulation signals can be obtained [16]. The selection of σ generates infinite number of modulators [14], [15].

$$V_{no} = \frac{V_{dc}}{2}(1 - 2\sigma) - V_{min}\sigma + V_{max}(\sigma - 1) \quad (9)$$

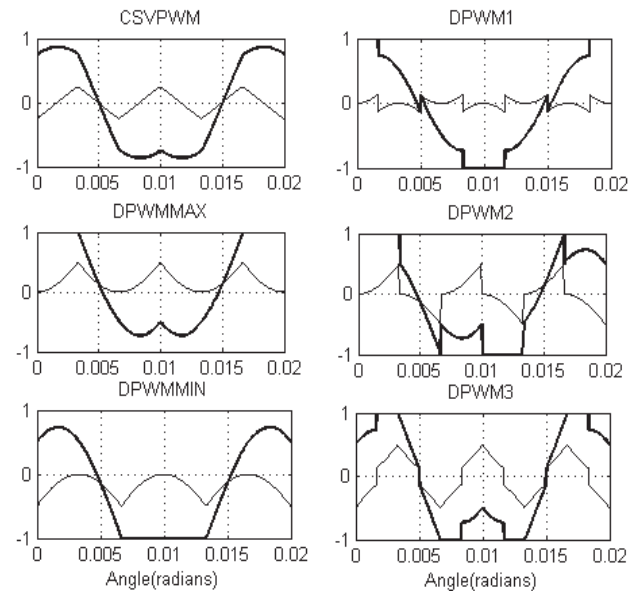


Figure 2. Modulation waveforms and their zero sequence signals of different DPWM methods

The generalized discontinuous modulation signal is given by (10),

$$\sigma = 1 - 0.5[1 + \text{sgn}(\cos 3(\omega t + \delta))] \quad (10)$$



where ω is the angular frequency of the reference voltage, δ is the modulation phase angle and $\text{sgn}(x)$ is the sign function defined as

$$\text{sgn}(x) = \begin{cases} +1, & x > 0 \\ 0, & x = 0 \\ -1, & x < 0 \end{cases} \quad (11)$$

For CSVPWM, σ is taken as 0.5. By varying the modulation angle δ , various DPWM methods can be generated. For example, $\delta = 0, \frac{-\pi}{6}, \frac{-\pi}{3}$ generates

DPWM1, DPWM2 and DPWM3 respectively which clamps for a period of 60° in every half cycle of the fundamental. The modulation waveforms and their zero sequence signals of some popular DPWM methods including CSVPWM methods are shown in Figure 2.

When $\sigma = 0$ and $\sigma = 1$, 120° clamping methods DPWMMAX and DPWMMIN are obtained. Thus by varying σ and δ the switching time periods of zero voltage vectors can be changed, so that different DPWM methods can be obtained.

Proposed DTC: The block diagram of the proposed GDPWM based DTC is shown in the Figure 3. With the proposed method ripples in torque and flux are reduced significantly and with fixed switching frequency. The proposed DTC retains all the advantages of the basic DTC, such as no coordinate transformation, robust to machine parameter variation, no current loop etc; the

error in the speed is fed to the PI controller which generates the reference value of the torque. Based on this reference value, the PI controller in turn generates the slip speed, when added with the actual rotor speed generates the speed of the reference stator flux vector $\overline{\psi_s^*}$.

For every sampling period $\overline{\psi_s}$ and the $\overline{\psi_s^*}$ are compared and the reference voltage space vectors are calculated as given by (12).

$$V_{ds}^* = R_s i_{ds} + \frac{\Delta \psi_{ds}}{T_s} \quad (12.1)$$

$$V_{qs}^* = R_s i_{qs} + \frac{\Delta \psi_{qs}}{T_s} \quad (12.2)$$

where,

$$\Delta \psi_{ds} = \psi_{ds}^* - \psi_{ds} \quad (13.1)$$

$$\Delta \psi_{qs} = \psi_{qs}^* - \psi_{qs} \quad (13.2)$$

The reference phase voltages are derived using the phase transformation from which imaginary switching times proportional to reference phase voltages are derived using (4). After obtaining the switching times for the active states and zero states the reference voltage vector is synthesized in an average sense with in a sub cycle. Thus using imaginary switching times space vector based GDPWM algorithm is implemented.

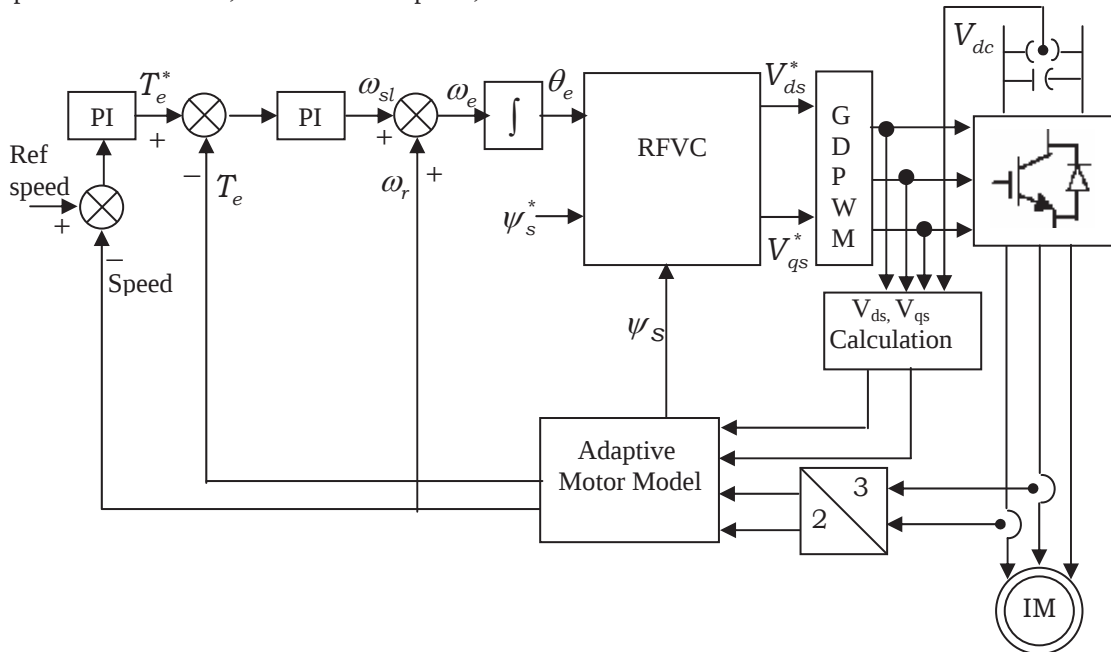


Figure.3. Block diagram of proposed DTC.

6. Simulation Results and Discussion

To verify the proposed scheme, numerical simulation has been carried out in Matlab/Simulink environment. Simulation is done using Runge-Kutta solver with a fixed step size of $10e-6$. The average switching frequency of the inverter is taken as 5kHz. The reference flux is taken as 1wb and starting torque is limited to 15 N-m. The induction motor used in the case study is a 1.5 KW, 1440 rpm, 4-pole, 3-phase induction motor having the

following parameters: $R_s = 7.83\Omega$, $R_r = 7.55\Omega$, $L_s = 0.4751H$, $L_r = 0.4751H$, $L_m = 0.4535H$, $J = 0.06 \text{ Kg.m}^2$. Simulation results for different DPWM based DTC including CDTC, showing the steady state plots are given in Figure.4 – Figure.9. Figure.4a shows the steady state ripples in stator currents, speed, torque and flux with the CDTC. It is observed that steady state ripples in torque and flux are high with the CDTC. Hence to reduce the steady state ripples and to get the constant switching



frequency with reduced switching losses a GDPWM algorithm using imaginary switching times is proposed. Figures 5-10 shows the simulation results obtained from the proposed method. Steady state ripples and its spectral performance is depicted from Figure 5-Figure 10. It can be observed that compared with the CDTC the performance in terms of ripple in torque and flux is significantly improved with the proposed DPWM methods. Also, observation of measured no load current waveforms and %THD's in line current, it can be concluded that performance of the different DPWM based DTC methods is superior to the CDTC.

Among the various DPWM methods DPWM2 is known for minimum switching losses. It is suitable for induction motor drives operating near 30° power factor angles [14]. Hence more elaborated simulation results showing different conditions of the drive, resembling transients during external load disturbances, starting transients, transients during step change in load, and transients during speed reversal are presented for DPWM2 based DTC, Figure 10. Figure 10b shows the transients in stator currents, speed, torque and flux during step change and non-linear variation in load. Figure 10c shows the starting transients under no-load condition. Figure 10d shows the transients during step change in load. Figure 10e shows the transients during speed reversal. Speed reversal command is given at 1.8sec to change the speed from +1300rpm to -1300rpm.

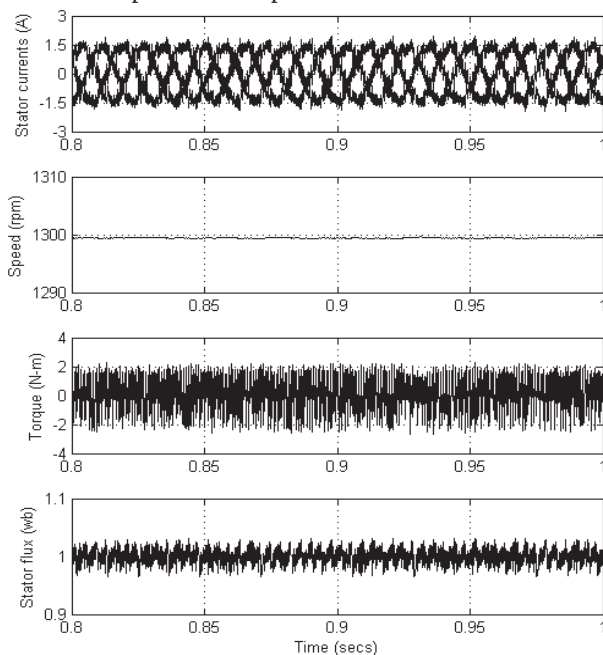


Figure 4a. CDTC: steady state plots (Reference speed=1300 rpm).
Fundamental (43.33Hz) = 1.39 , THD= 21.01%

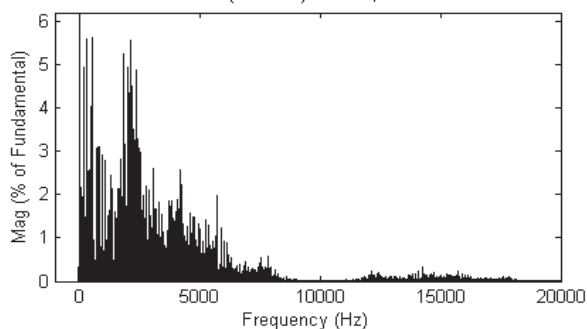


Figure 4b CDTC: Measured harmonic spectra

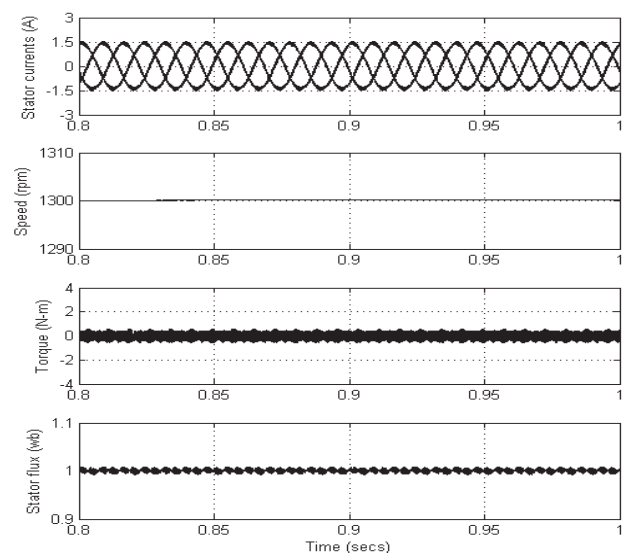


Figure 5a. DPWMMAX based DTC: steady state plots (at 1300 rpm)

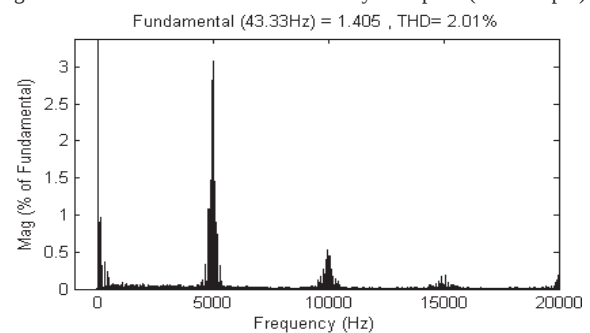


Figure 5b. DPWMMAX based DTC: Measured harmonic spectra

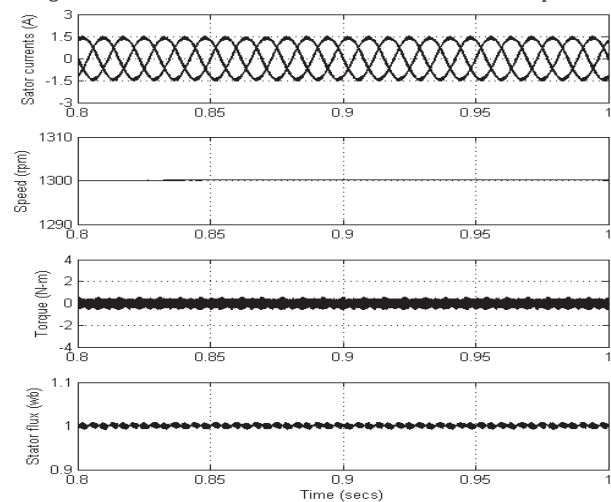


Figure 6a. DPWMMIN based DTC: steady state plots (at 1300 rpm)

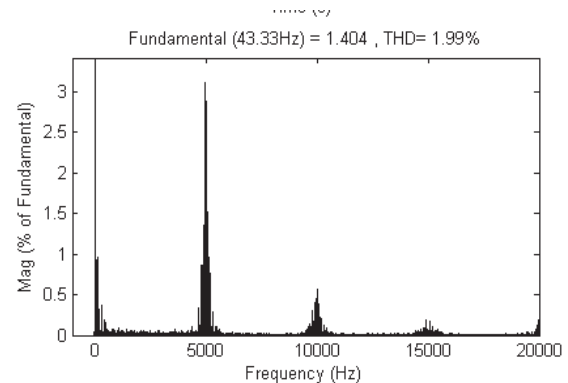


Figure 6b. DPWMMIN based DTC: Measured harmonic spectra



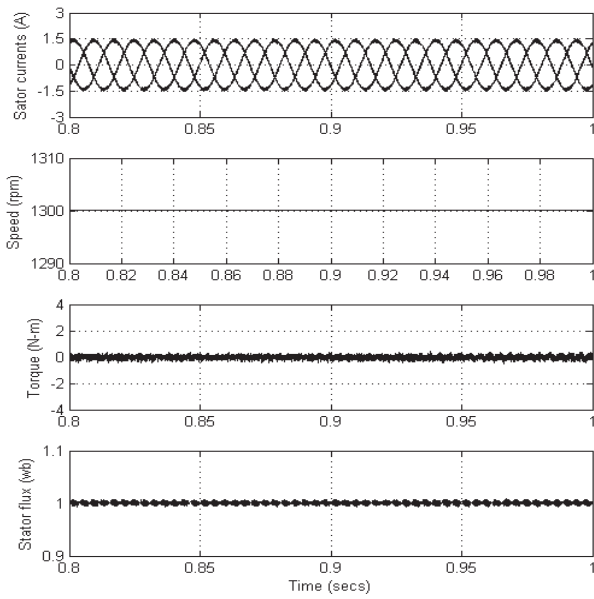


Figure.7a. DPWM1 based DTC: steady state plots (at 1300 rpm)

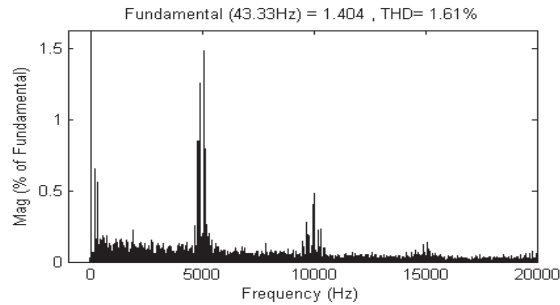


Figure.7b. DPWM1 based DTC: Measured harmonic spectra

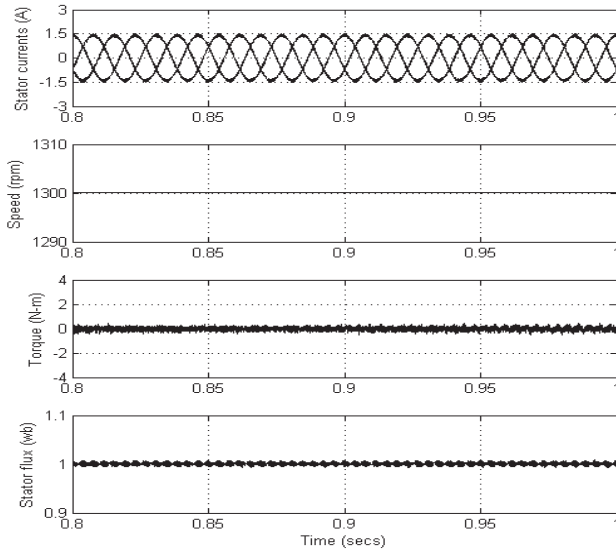


Figure.8a. DPWM2 based DTC: steady state plots (at 1300 rpm)

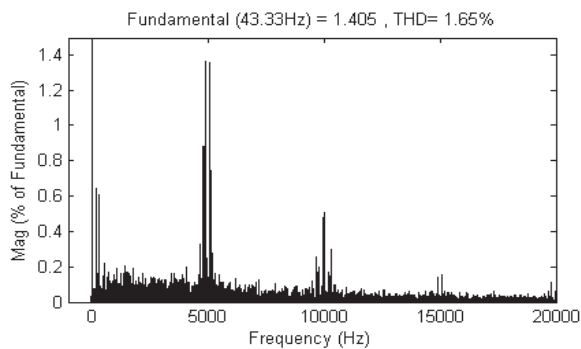


Figure.8b. DPWM2 based DTC: Measured harmonic

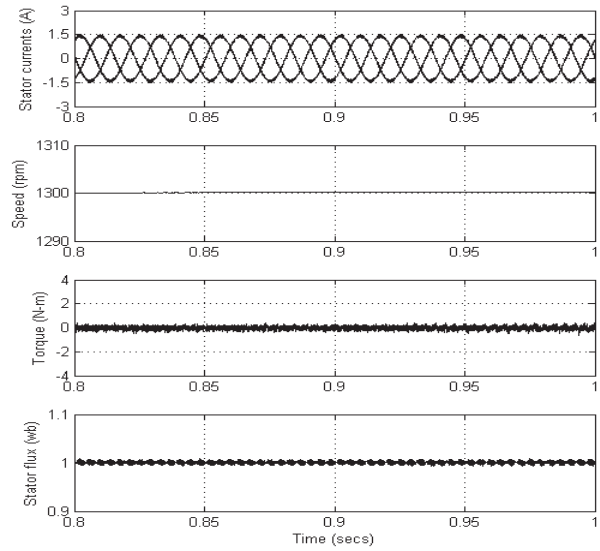


Figure 9a DPWM3 based DTC: steady state plots (at 1300 rpm)

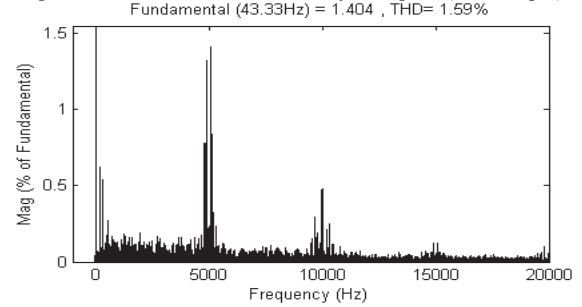


Figure.9b. DPWM3 based DTC: Measured harmonic spectra

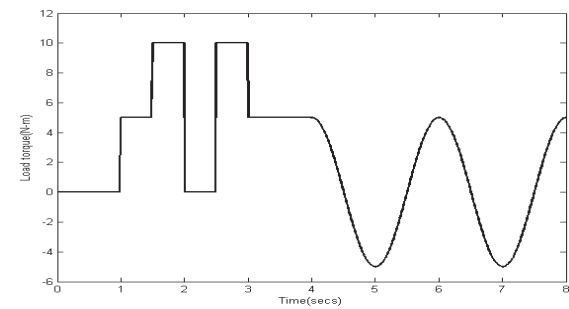


Figure.10a. External load torque disturbance.

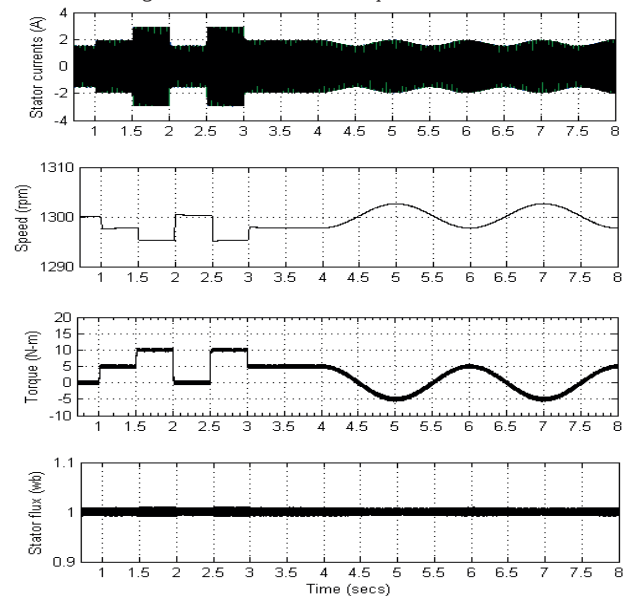


Figure.10b. DPWM2 based DTC: Transients showing the external load torque disturbance.



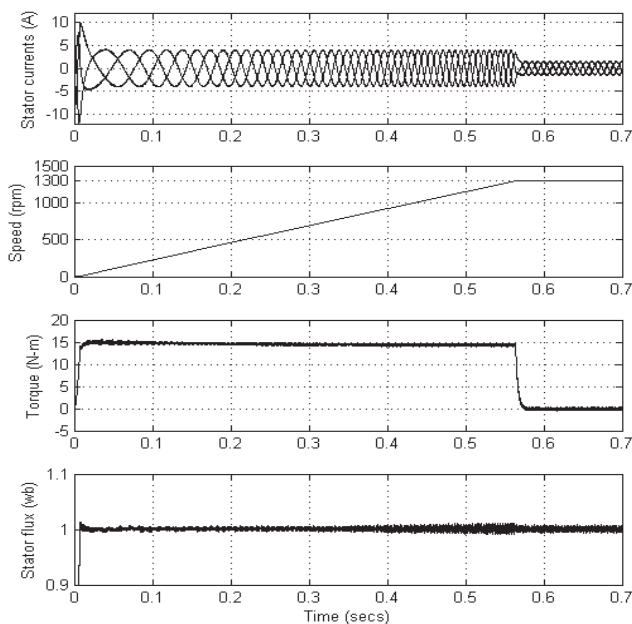


Figure 10c. DPWM2 based DTC: No load starting transients.

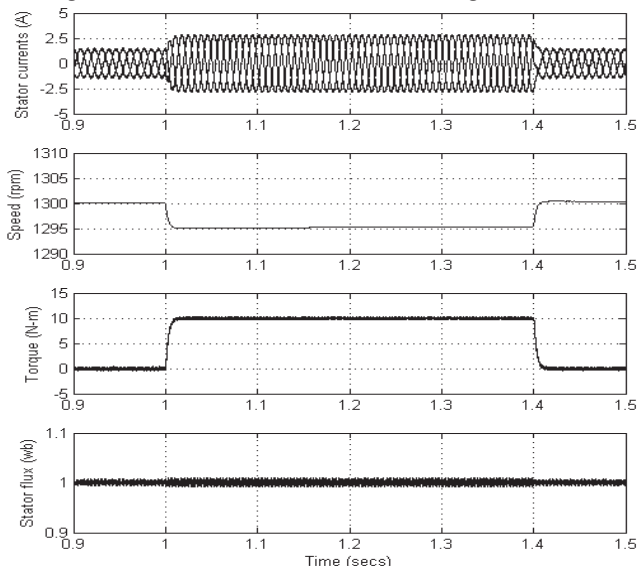


Figure 10d. DPWM2 based DTC: Transients during step change in load (A load of 10N-m is applied at 1sec and removed at 1.4 sec)

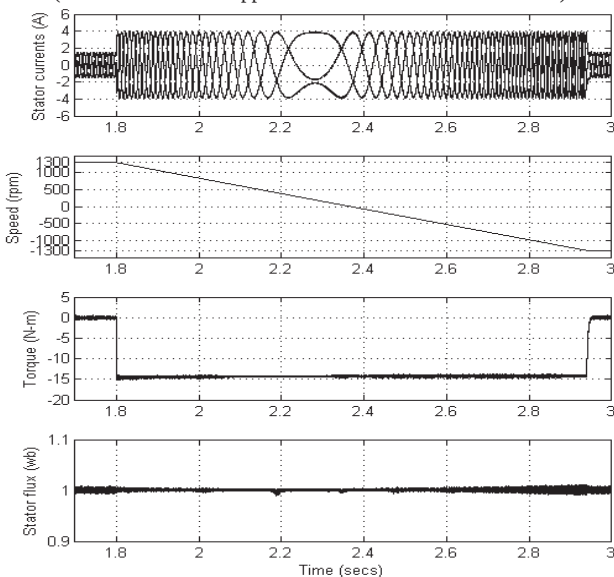


Figure 10e. DPWM2 based DTC: Transients during reversal of speed (Speed reversal command is given at 1.8 sec to change the speed from +1300 rpm to -1300 rpm).

7. Conclusion

Though CDTC is simple to implement, it has few drawbacks like significant steady state ripples in torque and flux, variable switching frequency because of the use of hysteresis controllers. To reduce these ripples and to get constant switching frequency SVPWM based DTC came into existence. Though SVPWM based DTC could solve most of the tribulations, deteriorated performance at high modulation indices became a significant setback. DPWM techniques give good performance in high modulation regions. Also switching losses of the inverter are relatively not as much of Continuous PWM techniques. In this paper to reduce the complexity involved in space vector based GDPWM the concept of imaginary switching times is implicated. With the proposed PWM algorithm, infinite DPWM methods can be developed. Also, it can be concluded that with the proposed algorithm in addition to the improved performance at high modulation indices, switching losses of the inverter and computational burden involved in conventional GDPWM algorithm are reduced significantly.

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Improved DTC relying on Hybrid Fuzzy-self tuning PI Regulator for the Permanent Magnet Synchronous Machine

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Abstract

We propose in this paper a Self-adaptation PI for speed regulation based on direct torque control (DTC) of Permanent Magnet Synchronous Machine (PMSM). Speed regulation with a conventional PI regulator reduces the speed control precision, increases the torque fluctuation, and consequentially causes low performances of the whole system. Using fuzzy logic method, by self-adaptation of conventional PI regulator parameters gives the appropriate Kp and Ki (proportional and integral coefficients respectively) which improve the system performance. Simulation results show that the ripples of both torque and flux are reduced remarkably, small overshooting and good dynamics of speed and torque.

Keywords: Permanent Magnet Synchronous Machine, Direct Torque Control, Fuzzy PI regulator.

1. Introduction

The PMSM control difficulty resides in the coupling of control variables such as flux and electromagnetic torque. Two principal strategies were developed almost at the same time in two different research centers, Direct Torque Control strategy was first introduced by I. Takahashi, in 1986 [1]. M. Depenbrock, develop a similar idea in 1988 under the name of Direct Self Control [2]. The DTC is one of the recent researched control schemes based on the decoupled control of stator flux and torque providing a quick and robust response with a simple implementation in AC drives. DTC has the advantages of simplicity, good dynamic performance, and insensitive to motor parameters except the stator resistance. In DTC strategy, the speed sensor is not essential for the flux and torque estimation. Direct Torque Control employs two hysteresis controllers to regulate the stator flux and torque, which results an approximate decoupling between the flux and the torque control. The key issue of DTC scheme is how to choose a suitable stator voltage vector to keep the stator flux and torque in their hysteresis band. The conventional DTC disadvantages are the high torque ripples and the slow transient response to the step changes in torque during

start-up. Several techniques developed to improve the DTC performance [3], [4], [5], [6], [7]. In paper [8] the authors replace the conventional PI by the fuzzy logic regulator when the output is the torque reference. In paper [9] fuzzy logic is used to replace the switching table of DTC which make possible to choose the very suitable voltage vector. The paper [6] realizes Fuzzy-PI speed regulation for induction machine; we utilize this idea for PMSM in this paper. The fuzzy control is nonlinear and adoptive in nature, giving robust performances in the face of parameter variations and load disturbance effects. The regulator inputs are speed error and its change. Self-adaptation PI regulator, based on conventional PI regulator; consist to adjust dynamically the conventional PI parameters kp and ki. Simulation results show that the method improves the Direct Torque Control system performances. This paper consists of mathematical model of PMS machine, direct torque control principle; fuzzy logic technique applied to DTC, simulation results with interpretation, and finally conclusion.

2. Motor equations in the α, β reference frame

In the stationary α - β reference frame, the model can be expressed as [8]:

$$u_{\alpha} = R_s \cdot i_{\alpha} + \frac{d\psi_{\alpha}}{dt} \quad (1)$$

$$u_{\beta} = R_s \cdot i_{\beta} + \frac{d\psi_{\beta}}{dt}$$

$$\begin{bmatrix} \psi_{\alpha} \\ \psi_{\beta} \end{bmatrix} = \begin{bmatrix} \cos \theta_r & -\sin \theta_r \\ \sin \theta_r & \cos \theta_r \end{bmatrix} \begin{bmatrix} \psi_d \\ \psi_q \end{bmatrix} \quad (2)$$

$$\psi_d = L_d \cdot i_d + \psi_m \quad (3)$$

$$\psi_q = L_q \cdot i_q$$

$$\begin{bmatrix} \psi_{\alpha} \\ \psi_{\beta} \end{bmatrix} = \begin{bmatrix} L_d \cdot i_{\alpha} \\ L_q \cdot i_{\beta} \end{bmatrix} + \psi_m \begin{bmatrix} \cos \theta_r \\ \sin \theta_r \end{bmatrix} \quad (4)$$

The mechanical equation given by:



$$\frac{d\omega_r}{dt} = \frac{p}{J} (T_{em} - T_L - B_m \omega_r) \quad (5)$$

In addition, the electromagnetic torque expressed as follow:

$$T_{em} = \frac{3}{2} p (\psi_{s\alpha} i_{s\beta} - \psi_{s\beta} i_{s\alpha}) \quad (6)$$

Where (u_{α}, u_{β}) , (i_{α}, i_{β}) and $(\psi_{s\alpha}, \psi_{s\beta})$ are the stator voltages, stator currents, and stator flux linkages in α, β reference frame, L_d, L_q are the d,q axes inductance, R_s is the stator resistance. ψ_m is the flux linkage of permanent magnet, p is the number of pole pairs, T_{em} is the electromagnetic torque, T_L is the load torque, B_m is the damping coefficient, ω_r is the rotor speed and J is the moment of inertia.

3. Direct torque control principles

3.1. Flux and torque estimation and control

In stationary reference frame, the machine stator voltage space vector is represented as follows:

$$V_s = R_s \cdot i_s + \frac{d\psi_s}{dt} \quad (7)$$

$$V_s = u_{s\alpha} + ju_{s\beta} = \sqrt{\frac{2}{3}} \left[V_{aN} + V_{bN} \cdot e^{j\frac{2\pi}{3}} + V_{cN} \cdot e^{j\frac{4\pi}{3}} \right] \quad (8)$$

Where: R_s , i_s , ψ_s stator resistance, current and flux respectively. V_{aN} , V_{bN} , V_{cN} the three phase voltage inverter outputs given as follows:

$$\begin{aligned} V_{aN} &= V_{sa} = \frac{U_c}{3} (2 \cdot C_1 - C_2 - C_3) \\ V_{bN} &= V_{sb} = \frac{U_c}{3} (2 \cdot C_2 - C_1 - C_3) \\ V_{cN} &= V_{sc} = \frac{U_c}{3} (2 \cdot C_3 - C_2 - C_1) \end{aligned} \quad (9)$$

U_c is the inverter DC supply voltage, C_1, C_2, C_3 are the switching table outputs, and they are relevant to the switching strategy.

From (7) we can estimate the stator flux as follow:

$$\psi_s = \psi_{s0} + \int (V_s - R_s \cdot i_s) \quad (10)$$

So we can write:

$$\begin{aligned} \psi_{s\alpha} &= \int (v_{s\alpha} - R_s i_{s\alpha}) dt \\ \psi_{s\beta} &= \int (v_{s\beta} - R_s i_{s\beta}) dt \end{aligned} \quad (11)$$

The module of the stator flux is:

$$\|\psi_s\| = \sqrt{\psi_{s\alpha}^2 + \psi_{s\beta}^2} \quad (12)$$

ψ_s is the stator flux vector, and ψ_{s0} is its initial value.

For simplicity, it is assumed the stator voltage drop $R_s i_s$ is small and neglected, the stator flux variation can be expressed as: $\Delta\psi_s \approx V_s \Delta t$.

We can estimate the electromagnetic torque using the following relation:

$$T_{em} = \frac{3}{2} p (\psi_{s\alpha} i_{s\beta} - \psi_{s\beta} i_{s\alpha}) \quad (13)$$

We can be controlled the change of torque by keeping the amplitude of the stator flux linkage constant and by controlling the rotating speed of the stator flux as fast as possible. We show in this section that both of the amplitude and rotating speed of the stator flux controlled by selecting the proper stator voltage vectors as shown in fig.1. The primary voltage vector V_s , is defined by the equation (8)

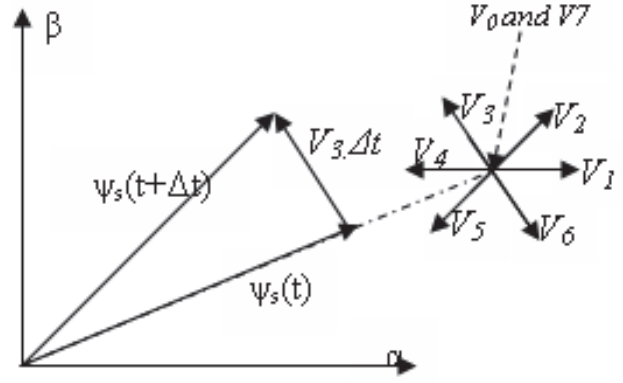


Figure1. Stator flux variation in stationary (α, β) frame

In the DTC technique, the inverter switches obtained using the flux and torque errors, and the position of the stator flux within the six-region control of the flux. The flux and torque errors evaluated as follows:

$$\Delta\psi = \psi_{sref} - \psi_s \quad (14)$$

$$\Delta T = T_{ref} - T_{em} \quad (15)$$

$$\theta = \tan^{-1} \left(\frac{\psi_{s\beta}}{\psi_{s\alpha}} \right) \quad (16)$$

Where θ is the angle between stator flux vector and α axis, $\psi_{s\alpha}, \psi_{s\beta}$ are the stator flux components in (α, β) reference frame.

3.2. Switching table:

In order to determine the inverter switching pattern using flux and torque errors, two hysteresis controllers are employing. The inverter is switched based on these errors and the position of the stator flux within the six-region control as can be seen from Table 1, in such a way that the inverter output voltage vector minimizes the flux and torque errors and determines the flux rotation direction.

n		1	2	3	4	5	6
Flux	Couple						
$K_{\psi}=1$	$K_{Cem}=1$	V_2	V_3	V_4	V_5	V_6	V_1
	$K_{Cem}=0$	V_7	V_0	V_7	V_0	V_7	V_0
	$K_{Cem}=-1$	V_6	V_1	V_2	V_3	V_4	V_5
$K_{\psi}=0$	$K_{Cem}=1$	V_3	V_4	V_5	V_6	V_1	V_2
	$K_{Cem}=0$	V_0	V_7	V_0	V_7	V_0	V_7
	$K_{Cem}=-1$	V_5	V_6	V_1	V_2	V_3	V_4

Table.1. DTC switching table.

4. Control structure

Figure (2) illustrates the PMSM drive scheme considered in this investigation. The drive consists of a Fuzzy PI speed controller, flux and torque controllers, space flux position, and PMSM. The rotor speed ω_r compared with the reference speed ω_{ref} . The resulting error and its rate of change are processed in the fuzzy PI speed controller for each sampling interval. The output of speed regulator considered as the reference torque T_{ref} .



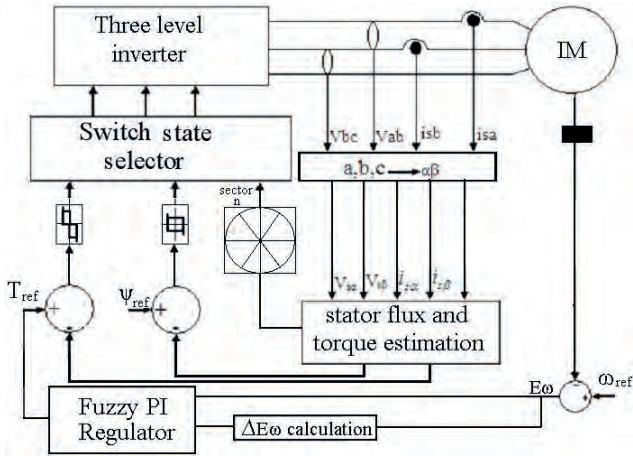


Figure2. A direct torque control scheme

5. Fuzzy controller

Speed error " $E\omega$," and its rate of change $\Delta E\omega$ are using as inputs to the fuzzy controller. Proportional coefficient Kp and integral coefficient Ki are the outputs of the fuzzy controller. The number of fuzzy segments is chosen to have maximum control with a minimum number of rules. Triangle and trapezoidal membership functions have been used. The fuzzy membership functions of input and output variables are shown in (Fig. 3) [6].

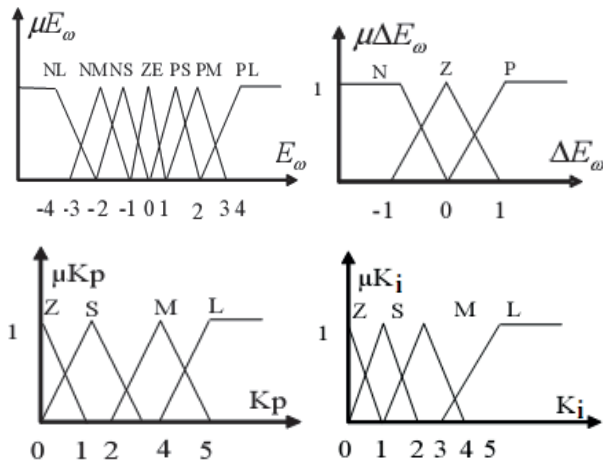


Figure3. The fuzzy membership functions of input and output variables.

From the experience of simulation and experiment, the range of coefficients k_p and k_i are [1, 5] and [0.005, 0.02], respectively [6]. The speed error universe of discourse is divided into seven fuzzy sets. {NL (negative large), NM (negative middle), NS (negative small), ZE (zero), PS (positive small), PM (positive middle), PL (positive large)}. rate of change $\Delta E\omega$ includes 3 fuzzy subsets, it is not necessary to subdivide it, because it's changes quickly in DTC. Output membership KP and KI , both contain four fuzzy subsets as shown in (fig. 3).

There are total of 21 rules as listed in table 2. Each control rule can be described using the inputs variables speed error " $E\omega$ ", it rate of change $\Delta E\omega$, and output variables (controller parameters k_i and k_p). The i^{th} rule R_i can be expressed as:

R_i : If $E\omega$ is A_i , $\Delta E\omega$ is B_i then KP is V_i , KI is W_i .

Where A_i , B_i , V_i and W_i denote the fuzzy subsets.

Kp, Ki		Eω						
		NL	NM	NS	ZE	PS	PM	PL
ΔEω	N	L, Z	M, S	S, M	M, L	S, M	M, S	L, Z
	Z	L, Z	M, S	L, M	Z, L	L, M	M, S	L, Z
	P	L, Z	M, M	L, L	Z, L	L, L	M, M	L, Z

Table. 2- Fuzzy Rules Base

The inference method used in this paper is Mamdani's procedure (inference) based on min-max decision. The firing strength (applied fuzzy operators) α_i for i^{th} rules is given by:

$$\alpha_i = \min(\mu_{A_i}(E\omega), \mu_{B_i}(\Delta E\omega)) \quad (17)$$

By fuzzy reasoning, Mamdani's minimum procedure gives:

$$\mu_{V_i}'(KP) = \min(\alpha_i, \mu_{V_i}(KP)) \quad (18)$$

$$\mu_{W_i}'(KI) = \min(\alpha_i, \mu_{W_i}(KI))$$

Where μ_A , μ_B , KP and KI are membership functions of sets A , B V and W of the variables $E\omega$, $\Delta E\omega$, Kp and Ki , respectively.

Thus, the membership functions μ_v and μ_w of the outputs KP and KI are given by:

$$\mu_v(KP) = \max_{i=1}^{21}(\mu_{V_i}'(KP)) \quad (19)$$

$$\mu_w(KI) = \max_{i=1}^{21}(\mu_{W_i}'(KI))$$

The Maximum criterion method is used for defuzzification. The final single-valued output is obtained by this method.

6. Simulation results

To verify the technique proposed, digital simulations based on MATLAB/SIMULINK have been implemented. The PMSM used for the simulations has the following parameters [10]:

Uc[V]	350
f[Hz]	50
Rs[Ω]	0.3
Ld[mH]	3.366
Lq[mH]	3.366
J [Kg/m ²]	10.8e-5
Bm[Nm/rad/s]	0
ψ _m [V/rad/s]	0.0776
p	5

Table3. PMS motor parameters

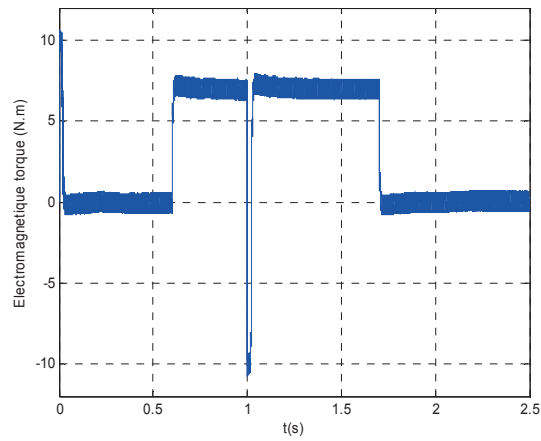
Conventional PI regulator coefficient:

$k_i = 0.0553$; $k_p = 1.441$.

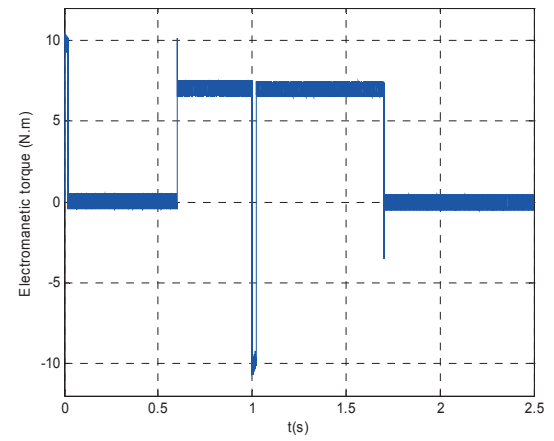
For proposed DTC and conventional DTC the dynamic responses of speed, flux, and torque for the starting



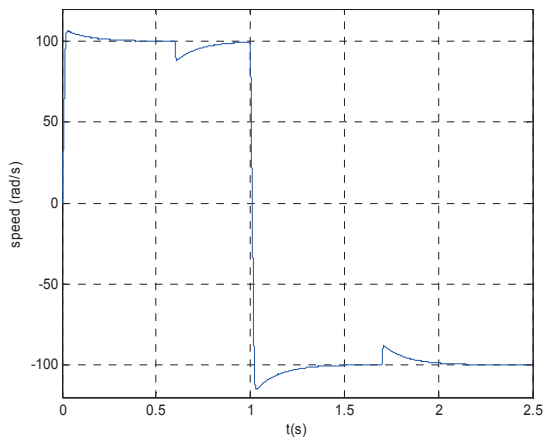
process without load $T_l=0$, we applied an load torque equal to $T_l=7\text{Nm}$ at 0.6s, at $t=1.7\text{s}$ we remove the load torque. We Inverse the speed at $t=1\text{s}$.



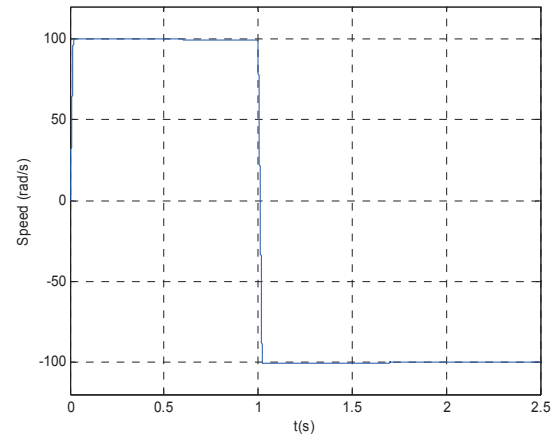
(a1)



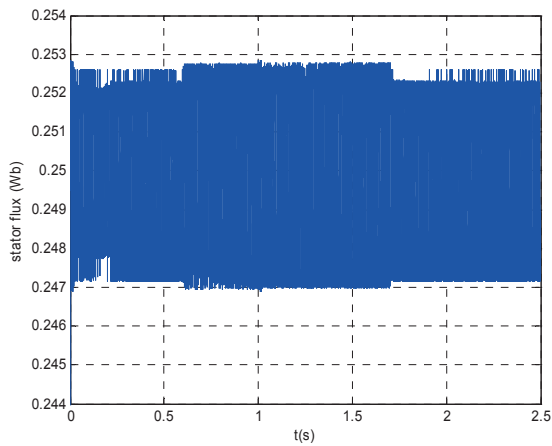
(a2)



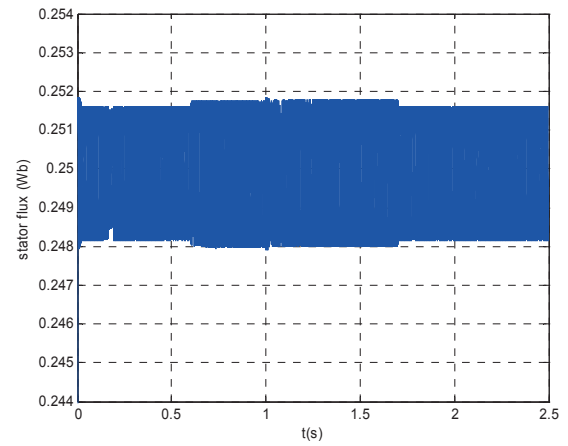
(b1)



(b2)



(c1)



(c2)

Figure4. (a1) Torque response, (b1) speed response, (c1) flux response, for DTC with conventional PI regulator.

Figure5. (a2) Torque response, (b2) speed response, (c2) flux response, for DTC with fuzzy PI regulation.



We compare between speed response for DTC with conventional regulator and fuzzy PI speed regulator.

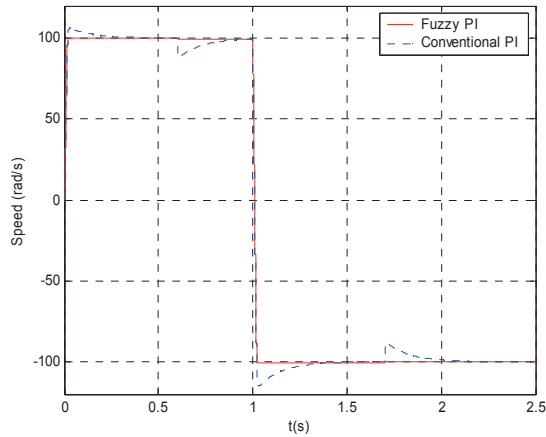


Figure6. Speed responses for DTC conventional PI regulator and DTC fuzzy PI regulator.

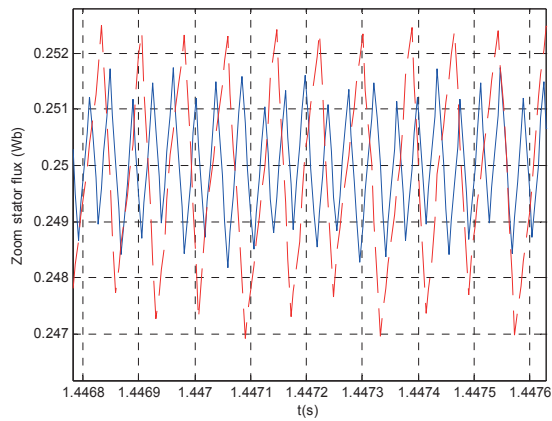


Figure7. Comparison between stator flux, dashed line stator flux using conventional PI regulator and solid line using fuzzy PI regulator.

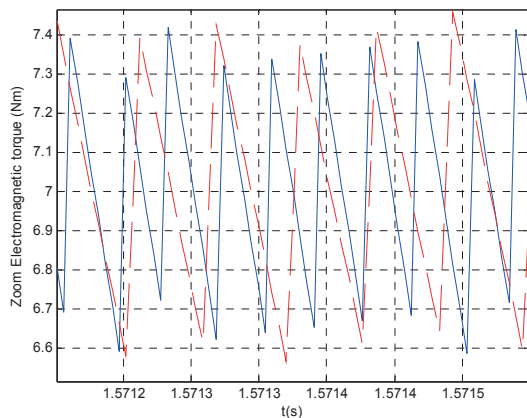
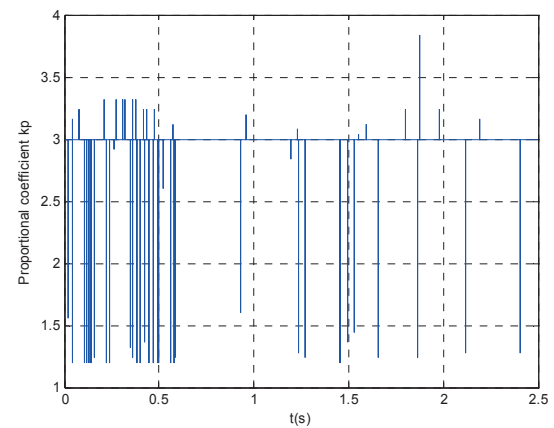
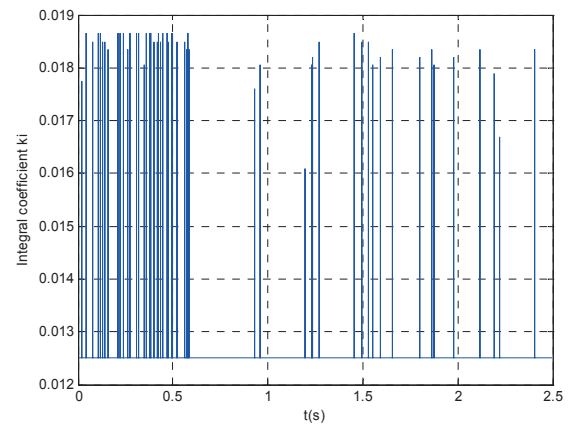


Figure8. Comparison between electromagnetic torque, dashed line stator flux using conventional PI regulator and solid line using fuzzy PI regulator.



(a)



(b)

Figure9. Proportional and integral coefficient estimated using fuzzy PI regulator.

The simulation results show that flux and torque responses are very fast for two DTC methods. By proposed DTC technique, the ripple of torque and flux in steady state is reduced remarkably compared with conventional DTC that reduce the acoustic noise and vibrations.

In fuzzy PI regulation good dynamic responses of torque with neglected influence of load disturbances in speed which restored its reference quickly.

Figures 7 (a) and (b) represent the estimated parameters k_p and k_i of the PI regulator, we observe that $k_p \in [1, 5]$ and $k_i \in [0.005, 0.02]$, as shown in section 5.

7. Conclusion

In this paper, a fuzzy logic direct torque control scheme using fuzzy PI regulator technique is presented. Using fuzzy logic technique, the k_p and k_i can be obtained dynamically that gives a fast speed response. The simulation results suggest that FLDTTC can achieve precise control of the stator flux and torque. Compared to conventional DTC, presented method the steady performances of ripples of both torque and flux are considerably improved.



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